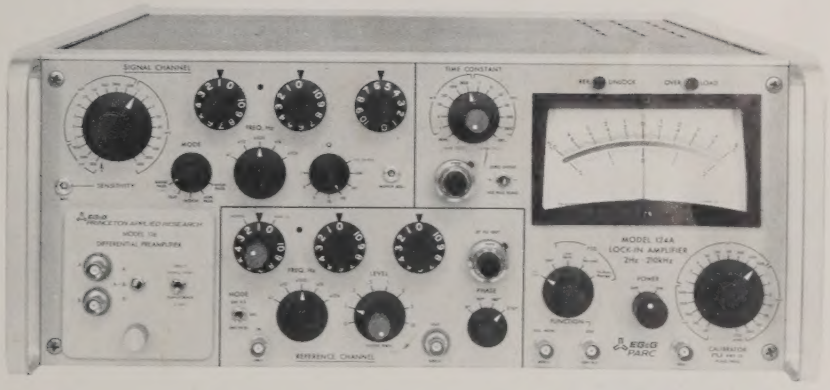
MODEL 124A
LOCK-IN AMPLIFIER

SEE SAFETY NOTICE
PRECEDING SECTION I
BEFORE OPERATING INSTRUMENT



OPERATING AND SERVICE MANUAL

MP MASTER
MA27841

EG&G
PRINCETON APPLIED RESEARCH

SHOULD YOUR EQUIPMENT REQUIRE SERVICE

A. Contact the factory (609/452-2111) or your local factory representative to discuss the problem. In many cases it will be possible to expedite servicing by localizing the problem to a particular plug-in circuit board.

B. If it is necessary to send any equipment back to the factory, we need the following information.

(1) Model number and serial number.

(2) Your name (instrument user).

(3) Your address.

(4) Address to which instrument should be returned.

(5) Your telephone number and extension.

(6) Symptoms (in detail, including control settings).

(7) Your purchase order number for repair charges (does not apply to repairs in warranty).

(8) Shipping instructions (if you wish to authorize shipment by any method other than normal surface transportation).

C. U.S. CUSTOMERS—Ship the equipment being returned to:

EG&G PRINCETON APPLIED RESEARCH
7 Roszel Road
(Off Alexander Road, East of Route 1)
Princeton, New Jersey

D. CUSTOMERS OUTSIDE OF U.S.A.—To avoid delay in customs clearance of equipment being returned, please contact the factory or the nearest factory distributor for complete shipping information.

E. Address correspondence to:

EG&G PRINCETON APPLIED RESEARCH
P. O. Box 2565
Princeton, NJ 08540

Phone: 609/452-2111
TELEX: 84 3409

WARRANTY

EG&G PRINCETON APPLIED RESEARCH warrants each instrument of its manufacture to be free from defects in material and workmanship. Obligations under this Warranty shall be limited to replacing, repairing or giving credit for the purchase price, at our option, of any instrument returned, freight prepaid, to our factory within ONE year of delivery to the original purchaser, provided prior authorization for such return has been given by our authorized representative.

This Warranty shall not apply to any instrument which our inspection shall disclose to our satisfaction, has become defective or unworkable due to abuse, mishandling, misuse, accident, alteration, negligence, improper installation or other causes beyond our control. Instruments manufactured by others, and included in or supplied with our equipment, are not covered by this Warranty but carry the original manufacturer's warranty which is extended to our customers and may be more restrictive. Certain subassemblies, accessories or components may be specifically excluded from this Warranty, in which case such exclusions are listed in the Instruction Manual supplied with each instrument.

We reserve the right to make changes in design at any time without incurring any obligation to install same on units previously purchased.

THERE ARE NO WARRANTIES WHICH EXTEND BEYOND THE DESCRIPTION HEREIN. THIS WARRANTY IS IN LIEU OF, AND EXCLUDES ANY AND ALL OTHER WARRANTIES OR REPRESENTATIONS, EXPRESSED, IMPLIED OR STATUTORY, INCLUDING MERCHANTABILITY AND FITNESS, AS WELL AS ANY AND ALL OTHER OBLIGATIONS OR LIABILITIES OF EG&G PRINCETON APPLIED RESEARCH, INCLUDING, BUT NOT LIMITED TO, SPECIAL OR CONSEQUENTIAL DAMAGES. NO PERSON, FIRM OR CORPORATION IS AUTHORIZED TO ASSUME FOR EG&G PRINCETON APPLIED RESEARCH ANY ADDITIONAL OBLIGATION OR LIABILITY NOT EXPRESSLY PROVIDED FOR HEREIN EXCEPT IN WRITING DULY EXECUTED BY AN OFFICER OF EG&G PRINCETON APPLIED RESEARCH.

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SAFETY NOTICE
(Read Before Operating Instrument)

A. INTRODUCTION

The Model 124A to which this instruction manual applies has been supplied in a safe condition. This manual contains some information and warnings that have to be followed by the user to ensure safe operation and to retain the instrument in a safe condition.

The described apparatus has been designed for indoor use.

B. INSPECTION

Newly received apparatus should be inspected for shipping damage. If any is noted, notify EG&G PARC and file a claim with the carrier. Be sure to save the shipping container for inspection by the carrier.

WARNING!

THE PROTECTIVE GROUNDING COULD BE RENDERED INEFFECTIVE IN DAMAGED APPARATUS. DAMAGED APPARATUS SHOULD NOT BE OPERATED UNTIL ITS SAFETY HAS BEEN VERIFIED BY QUALIFIED SERVICE PERSONNEL. DAMAGED APPARATUS WAITING FOR SAFETY VERIFICATION SHOULD BE TAGGED TO INDICATE TO A POTENTIAL USER THAT IT MAY BE UNSAFE AND THAT IT SHOULD NOT BE OPERATED.

C. SAFETY MECHANISM

As defined in IEC Publication 348 (*Safety Requirements for Electronic Measuring Apparatus*), the Model 124A is Class I apparatus, that is, apparatus that depends on connection to a protective conductor to earth ground for equipment and operator safety. Before any other connection is made to the apparatus, the protective connection is made via the earth ground prong of the power cord plug. The power cord plug shall only be inserted in a socket outlet provided with the required earth ground contact. The protective action must not be negated by the use of an extension cord without a protective conductor, by use of an "adapter" that doesn't maintain earth ground continuity, or by any other means.

WARNING!

ANY INTERRUPTION OF THE PROTECTIVE CONDUCTOR INSIDE OR OUTSIDE THE APPARATUS OR DISCONNECTION OF THE PROTECTIVE EARTH TERMINAL MAY MAKE THE APPARATUS DANGEROUS. INTENTIONAL INTERRUPTION IS PROHIBITED.

The power cord plug provided is of the type illustrated in Figure 1. If the provided plug is not compatible with the available power sockets, the plug or power cord should be replaced with an approved type of compatible design.

WARNING!

IF IT IS NECESSARY TO REPLACE THE POWER CORD OR THE POWER CORD PLUG, THE REPLACEMENT CORD OR PLUG MUST HAVE THE SAME POLARITY AS THE ORIGINAL. OTHERWISE A SAFETY HAZARD FROM ELECTRICAL SHOCK, WHICH COULD RESULT IN INJURY OR DEATH, MIGHT RESULT.

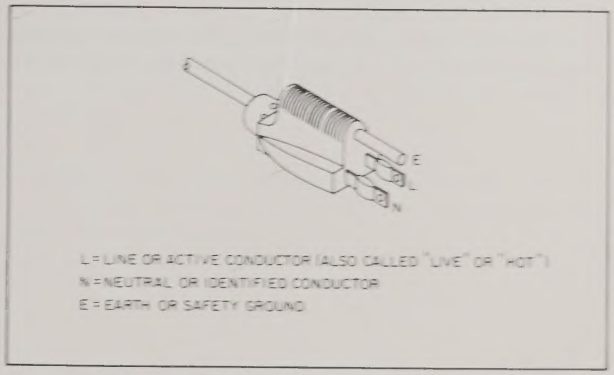


Figure 1. POWER CORD PLUG WITH POLARITY INDICATIONS

D. POWER VOLTAGE SELECTION

Before plugging in the power cord, make sure that the instrument is set for operation from the voltage of the ac power supply.

CAUTION!

THE APPARATUS DESCRIBED IN THIS MANUAL MAY BE DAMAGED IF IT IS SET FOR OPERATION FROM 110 V AC AND TURNED ON WITH 220 V AC APPLIED TO THE POWER INPUT CONNECTOR.

The line voltage is selected by means of a rear-panel switch. Depending on the switch position, either "115" or "230" (both are printed on the switch) will be visible to the operator. For operation from a line voltage of nominally 115 V, position the switch so that "115" shows. For operation from a line voltage of nominally 230 V, position the switch so that "230" shows.

E. LINE FUSE

The ac line fuse holder is located at the rear panel near the line cord. For operation from a line voltage of nominally 115 V, the fuse should be a slow-blow type rated at 1/2 A. For operation from a line voltage of nominally 230 V, the fuse should be a slow-blow type rated at 1/4 A. The voltage rating of the fuse should be 125 V or higher for 115 V operation and 250 V or higher for 230 V operation. Only a LITTELFUSE type 3AG or equivalent fuse should be used.

WARNING!

TO AVOID THE POSSIBILITY OF A SAFETY HAZARD FROM ELECTRICAL SHOCK WHICH COULD RESULT IN INJURY OR DEATH, DISCONNECT THE POWER CORD BEFORE REMOVING OR INSTALLING A FUSE. FUSES SHOULD ONLY BE CHANGED BY QUALIFIED SERVICE PERSONNEL WHO ARE AWARE OF THE HAZARD INVOLVED.

F. VENTILATION

The Model 124A does not have any special ventilation requirements. It can be operated on a bench or in a rack, the only constraint being that the ambient temperature be in the range of 15°C to 45°C.

G. DEFECTS AND ABNORMAL STRESSES

Whenever it is likely that the protection provided by the connection to earth ground has been impaired, the apparatus shall be made inoperative and secured against any unintended operation. The protection is likely to be impaired if, for example, the apparatus:

(1) Shows visible damage,

(2) Fails to perform the intended measurements,

(3) Has been subjected to prolonged storage under unfavorable conditions,

(4) Has been subjected to severe transport stresses.

Such apparatus should not be used until its safety has been verified by qualified service personnel.

SECTION I
CHARACTERISTICS

1.1 INTRODUCTION

The Model 124A Lock-In Amplifier accurately measures the rms amplitude and phase of weak signals buried in noise. Signals in the range of picovolts up to 500 millivolts at frequencies from 0.2 Hz to 210 kHz can be measured quickly and precisely. Meter and voltage outputs are provided for the amplitude, and the phase of the signal may be read from a dial. These measurements are with reference to a synchronizing signal supplied to, or supplied by, the Model 124A. In either the External or External f/2 mode of operation, the instrument will accept any reference waveform that crosses its mean twice each cycle, and will lock to and track that signal over a 100:1 frequency range. In the Internal mode, the frequency is determined by front-panel dials or by an externally derived voltage.

A selection of plug-in preamplifiers is available for providing optimum low-noise performance over a wide range of input frequencies and source resistances. After preamplification, noise and harmonics accompanying the signal are attenuated in the Signal Channel by filtering out all frequencies except the band in which the signal lies. Flat, band pass, band reject, high pass, and low pass filtering modes may be selected. The remaining band of frequencies is converted to an equivalent bandwidth about dc by a synchronous detector, which is locked to the synchronizing signal. A low-pass filter eliminates frequency components above dc, so that the detector output is a dc voltage proportional to the in-phase component of the fundamental signal. Proper selection of signal channel and output channel filtering parameters can render the final noise bandwidth extremely narrow. The rms value of the fundamental signal is indicated on the panel meter when the synchronous detection phase is adjusted for maximum detector output.

A switch is provided that allows drift to be traded for dynamic reserve. In addition, an output dc offset feature is provided to allow higher sensitivity settings for relatively steady signals. These features permit selection of the optimum operating mode for each experimental situation.

Other design features include selection of output filter time constants to 300 seconds, optional digital panel meter with BCD output, a built-in calibrator, and independent use of the phase-lockable oscillator and tuned amplifier for general-purpose laboratory work. The Model 124A may also be used as a conventional wideband laboratory voltmeter.

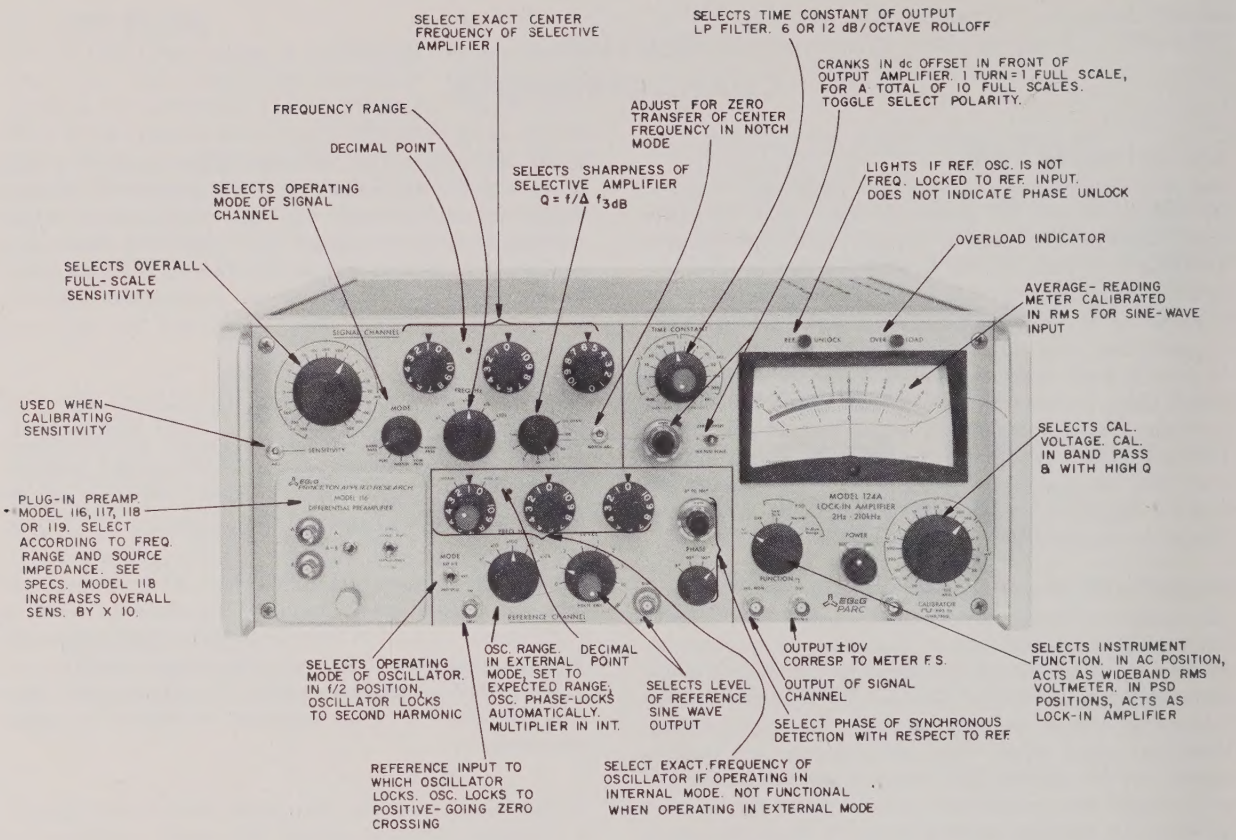
Accessories include an ac zero offset, several light choppers, a computer interface system, and a wide assortment of low-noise preamplifiers.

Specification	Model 116	Model 117	Model 118	Model 119
Input Z	Selected by front panel switch: Direct: 100 meg. SE/DE Transformer[a]: Low Z SE/DE	100 megohms SE/DE	10 kilohms SE/DE	Selected by front panel switch: Direct: 100 meg. SE/DE Transformer: Low Z SE/DE
Bandwidth	Direct: 0.2 Hz – 210 kHz Transformer[b]: 1.5 Hz-10 kHz	0.2 Hz – 210 kHz	0.2 Hz – 210 kHz	Direct: 0.2 Hz – 210 kHz Transformer[b]: 1 kHz-210 kHz
Common Mode Rejection Ratio	Direct: 120 dB at 60 Hz Transformer: 140 dB at 60 Hz	120 dB at 60 Hz	110 dB at 60 Hz	Direct: 120 dB at 60 Hz Transformer: 120 dB at 60 Hz
Full Scale Sensitivity	Direct: 100 nV Transformer: 1 nV[a]	100 nV	10 nV	Direct: 100 nV Transformer: 1 nV
Maximum Input Voltage	Direct: ±200 V dc Transformer: 10 mV rms sine wave	±200 V dc	±5 V	Direct: ±200 V dc Transformer: 10 mV rms sine wave

[a]May be wired for 1:50 to 1:350 turns ratio. Standard is 1:100.
[b]Varies with source impedance.

NOTE: A current-sensitive preamplifier, Model 184, is also available. See ACCESSORIES list at end of specs.

Table I-1. PREAMPLIFIER SPECIFICATIONS


SELECT EXACT CENTER FREQUENCY OF SELECTIVE AMPLIFIER
SELECTS TIME CONSTANT OF OUTPUT LP FILTER. 6 OR 12 dB/OCTAVE ROLLOFF
CRANKS IN dc OFFSET IN FRONT OF OUTPUT AMPLIFIER. 1 TURN = 1 FULL SCALE, FOR A TOTAL OF 10 FULL SCALES. TOGGLE SELECT POLARITY.
FREQUENCY RANGE
ADJUST FOR ZERO TRANSFER OF CENTER FREQUENCY IN NOTCH MODE
DECIMAL POINT
SELECTS SHARPNESS OF SELECTIVE AMPLIFIER Q = f/Δ f3dB
LIGHTS IF REF. OSC. IS NOT FREQ. LOCKED TO REF. INPUT. DOES NOT INDICATE PHASE UNLOCK
SELECTS OPERATING MODE OF SIGNAL CHANNEL
OVERLOAD INDICATOR
SELECTS OVERALL FULL-SCALE SENSITIVITY
AVERAGE-READING METER CALIBRATED IN RMS FOR SINE-WAVE INPUT
USED WHEN CALIBRATING SENSITIVITY
SELECTS CAL. VOLTAGE. CAL. IN BAND PASS & WITH HIGH Q
PLUG-IN PREAMP. MODEL 116, 117, 118 OR 119. SELECT ACCORDING TO FREQ. RANGE AND SOURCE IMPEDANCE. SEE SPECS. MODEL 118 INCREASES OVERALL SENS. BY X 10.
SELECTS OPERATING MODE OF OSCILLATOR. IN f/2 POSITION, OSCILLATOR LOCKS TO SECOND HARMONIC
OSC. RANGE. DECIMAL IN EXTERNAL POINT MODE, SET TO EXPECTED RANGE, OSC. PHASE-LOCKS AUTOMATICALLY. MULTIPLIER IN INT.
SELECTS LEVEL OF REFERENCE SINE WAVE OUTPUT
OUTPUT ±10V CORRESP. TO METER F.S.
OUTPUT OF SIGNAL CHANNEL
SELECTS INSTRUMENT FUNCTION. IN AC POSITION, ACTS AS WIDEBAND RMS VOLTMETER. IN PSD POSITIONS, ACTS AS LOCK-IN AMPLIFIER
SELECT PHASE OF SYNCHRONOUS DETECTION WITH RESPECT TO REF.
REFERENCE INPUT TO WHICH OSCILLATOR LOCKS. OSC. LOCKS TO POSITIVE-GOING ZERO CROSSING
SELECT EXACT FREQUENCY OF OSCILLATOR IF OPERATING IN INTERNAL MODE. NOT FUNCTIONAL WHEN OPERATING IN EXTERNAL MODE

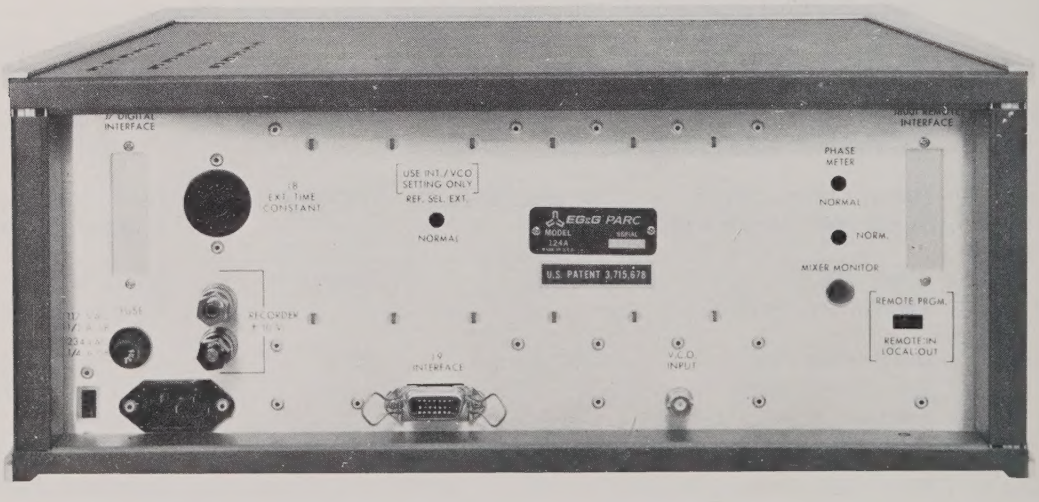


Figure I-1. MODEL 124A LOCK-IN AMPLIFIER

1.2 SPECIFICATIONS

1.2A SIGNAL CHANNEL SPECIFICATIONS

Frequency Range
Model 124A: 2 Hz – 210 kHz
Model 124AL: 0.2 Hz – 210 kHz*

Sensitivity: 21 full-scale ranges in 1-2-5 sequence. Full-scale voltages are determined by the choice of preamplifier. Sensitivity and all other preamplifier-determined specifications are given in Table I-1.**

Signal Channel Modes of Operation

(1) FLAT: Flat response within ±1% from 10 Hz to 110 kHz, ±2% from 110 kHz to 210 kHz, and ±10% below 10 Hz.

(2) BANDPASS: Provides a tunable bandpass response with the center frequency set by front-panel digital dials over a range of 2 Hz to 110 kHz. Setting accuracy is within ±2% or 0.05 Hz, whichever is greater. Bandwidth is adjustable over a range of 1% to 100% (at 3 dB points), corresponding to a range of Q between 100 and 1, by means of the front-panel Q control.

(3) NOTCH: Essentially the same as the Flat mode, but with the addition of a tunable notch that provides up to 80 dB of attenuation at any specific frequency. The notch is tuned with the same controls as set the bandpass frequency.

(4) LOW PASS: Essentially the same as the Flat mode, but with the addition of a low pass filter that provides a 12 dB per octave rolloff above the set frequency.

(5) HIGH PASS: Essentially the same as the Low Pass mode, but with the substitution of a high pass filter in place of the low pass filter.

1.2B REFERENCE CHANNEL SPECIFICATIONS

Modes

(1) INTERNAL: Frequency of the internal reference oscillator is set by means of front-panel digital dials and/or rear-panel VCO control voltage. Setting accuracy is within ±2% or 0.05 Hz, whichever is greater. VCO control voltage of 0 to ±10 V corresponds to the full frequency range on all bands. VCO input impedance is 10 kilohms. The amplitude stability is typically 0.01%. The frequency stability is typically 0.05% of the set frequency.

(2) EXTERNAL: The internal reference oscillator will lock in both frequency and phase to virtually any externally generated signal crossing its mean only twice each cycle. Maximum input voltage is ±20 V dc. Minimum time required on either side of the mean is 100 ns. Amplitude excursion must be at least 50 mV above and below the mean. Input impedance is 1 megohm.

When locked on, the reference oscillator will track the external signal over a frequency range of 100:1 within the range of the set band of frequencies. Maximum frequency acquisition (lock-on) times for each frequency band are given in the following table.

BAND	FREQUENCY RANGE	MAXIMUM TIME*
X1	0.2 Hz to 21 Hz	15 minutes
X10	2 Hz to 210 Hz	2 minutes
X100	20 Hz to 2.1 kHz	10 seconds
X1 K	200 Hz to 21 kHz	2 seconds
X10 K	2.1 kHz to 210 kHz	2 seconds

Once the frequency has locked, the phase will track at the rate shown in the diagram on the following page.

Phase Adjustment: Calibrated 10 turn potentiometer provides 0-100° phase shift. Linearity of phase setting is within ±2° from 2 Hz to 21 kHz, and within ±5° from 21 kHz to 210 kHz. Resolution is 0.1°. A four-position quadrant switch provides 90° phase shift increments.

1.2C DEMODULATOR CHARACTERISTICS

ACVM: An ACVM position on the function switch permits the Model 124A to be used as a conventional or frequency-selective ac voltmeter. Accuracy is within ±1% from 2 Hz to 20 kHz, increasing to ±10% at 210 kHz.

Dc Output Stability and Noise: Dependent on the operating mode selected by the front-panel Function switch, as shown in the Stability & Noise Table (next page).

*Model 124AL has significantly longer severe-overload recovery time (80 s vs 30 s with 5000 times full scale overload for one minute).

**Two additional preamplifiers, the Model 184 and the Model 185, are available. For information, contact the factory or the factory representative in your area.

*Time can be shortened appreciably by momentarily switching to Internal Mode and manually setting the oscillator to the proper frequency.

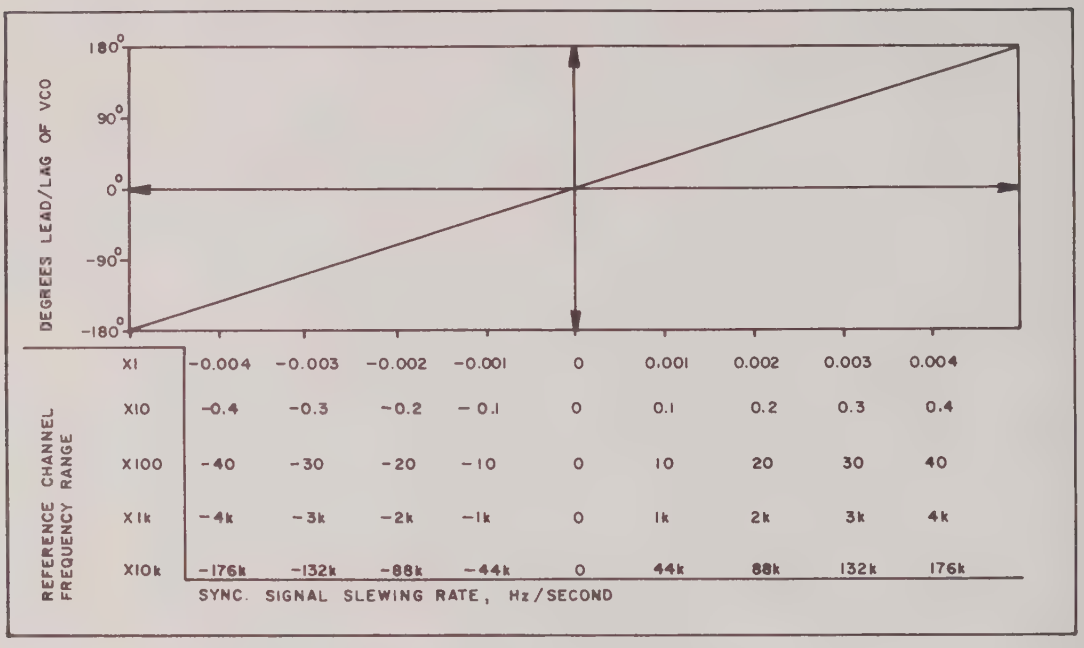


Figure 1-2. SYNC. SIGNAL SLEWING RATE

Function	Output Stability	Output Noise*
LOW DRIFT	<150 μV/°C	<100 μV rms
NORMAL	<1 mV/°C	<1 mV rms
HIGH DYN RANGE	<10 mV/°C	<10 mV rms

Demodulator Overload Limits: Dependent on Function switch setting as follows (see Subsection 3.2I for over-ride considerations).

LOW DRIFT ..... 10 x full scale
NORMAL ..... 100 x full scale
HIGH DYN RANGE ..... 1000 x full scale

This limit is defined as the ratio, at the input, of the maximum pk-pk voltage of a non-coherent signal, before overload, to the pk-pk voltage of a full-scale coherent sinewave. Note that, in terms of pk-pk noise to rms signal, the instrument will accept, without overload, interfering signals having an amplitude up to 3000 times the sensitivity setting. (See discussion in Subsection 3.2I.)

Filter Time Constants: 1 ms to 300 s in 1-3-10 sequence, and a minimum time constant position having a time constant of less than 1 ms (determined by internal stray capacitance). The External position allows capacitance to be added via a rear-panel connector to obtain special values of time constant. Either 6 or 12 dB/octave rolloff as selected by means of front-panel switch is provided.

*Measured with time constant of 1 s and 12 dB/octave rolloff.

Equivalent Noise Bandwidth: 416 μHz minimum (300 s time constant with 12 dB/octave rolloff).

Zero Suppress: Calibrated control permits off-setting zero by ±1000% of full scale on Normal and High Dynamic Range only.

System Gain Stability: 100 ppm/°C; 100 ppm/24 Hr in the Flat mode and with Function switch set to NORMAL.

1.2D OUTPUTS

Meter Reading: Choice of either center-zero or lefthand-zero panel meter of taut-band construction, providing 0.5% linearity.

Optional Digital Readout: The Model 124A may be ordered with an optional digital readout in place of the standard panel meter. The readout is a 3½ digit display with a linearity of 0.05% of the reading, ±1 count. In addition, a BCD output is provided at the rear panel. The output levels are DTL/TTL compatible: Logic 0 = +0.2 V ±0.2 V, 5 mA maximum sinking current; Logic 1 = +3.5 V ±1.0 V, 100 μA maximum sourcing current.

Function Out: A dc signal corresponding to the panel-meter reading. An output of 10 V corresponds to full-scale deflection. The output impedance is 1 kΩ.

Signal Monitor: Enables continuous monitoring of the signal channel output ahead of the demodulator. In LO DRIFT operation, a full scale rms input sinewave gives a 100 mV rms sinewave at the Signal Monitor jack. In NORMAL operation, the signal monitor output with a full-scale input is 10 mV, and in HI, it is 1 mV. Dynamic

Operating Dynamic Range Tradeoff	Output Dynamic Range	PSD Dynamic Reserve	PSD Dynamic Range	Total Dynamic Range
LO DRIFT	6.6×10^4	10	6.6×10^5	6.6×10^8
NORMAL	10^4	10^2	10^6	10^9
HI DYN RNG (Reserve)	10^3	10^3	10^6	10^8

Table I-2. MODEL 124A DYNAMIC RANGE SPECIFICATIONS

Over-ride considerations apply as explained in Subsection 3.2I. Output impedance is 600 Ω.

Internal line frequency pickup is less than 20 nV rms (referred to the Direct inputs of a Type 116 Preamplifier) in any Signal Channel mode except Bandpass and Notch, where the level may rise to 500 nV at highest Q settings.

Reference Channel: A sinewave output at the reference oscillator frequency. Amplitude is continuously adjustable by means of the front-panel Level control over a range of 0 V to 10 V rms with less than 2% distortion. Output impedance is 600 ohms.

1.2E DYNAMIC RANGE SPECIFICATIONS

Vary as a function of the operating Dynamic Range Tradeoff as indicated in the table above.

1.2F OTHER CHARACTERISTICS

Overload: Front-panel light indicates overload at critical circuits.

Reference Unlock: Front-panel light indicates that the reference oscillator has not completed frequency lock.

Internal Calibrator: Square-wave calibrator signal supplied. Rms amplitude of fundamental frequency component adjustable from 20 nV to 100 mV in 1-2-5 sequence. Typical accuracy indicated in Figure I-3.

Ambient Temperature Range: Unit can be operated at ambient temperatures ranging from 15°C to 45°C.

Auxiliary Power Output: Regulated ±24 V at up to 100 mA is available at rear-panel connector.

Power Requirements: 105-125 or 210-250 V; 50-60 Hz; unit can also be powered from batteries by supplying ±31 V to rear panel connector. Batteries must be able to supply at least 400 mA at +31 V and 360 mA at −31 V.

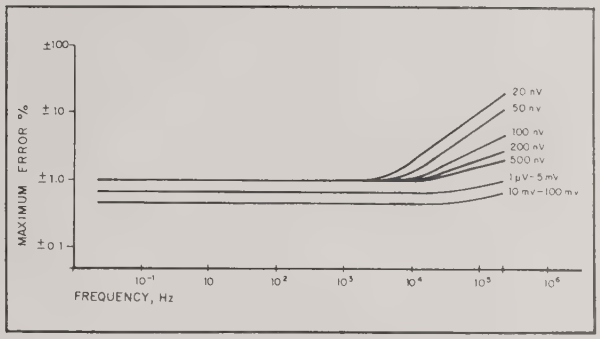


Figure I-3. TYPICAL CALIBRATOR ACCURACY

Size: 17-1/8" W x 7" H x 18-1/4" D (43.6 cm W x 17.8 cm H x 46.5 cm D).

Weight: 34 lbs (15.5 kg).

Accessories: Model 123 AC Zero Offset provides square wave at the reference frequency which can be used to suppress signals at the input of the Model 124A. Other accessories include a computer interface system, fixed and variable speed light choppers, and a broad selection of special purpose preamplifiers. The AM-1, AM-2, and 190 input transformers allow better noise performance to be achieved when using a high-input impedance preamplifier to process a signal arising in a low source impedance.

The Model 184 Current-Sensitive Preamplifier is also available. This preamplifier, which plugs in like the Models 116, 117, 118, and 119 Preamplifiers, provides 1 V out for input currents ranging from 1 nA to 10 μA as selected by a front-panel Range switch. Frequency range varies with sensitivity, being 2 Hz to 3 kHz on the 1 nA range and 2 Hz to 200 kHz on the 10 μA range.

*NOTE: These are maximum values and do not apply for all positions of the Sensitivity switch. For a general discussion of the meaning of these terms and their significance, see Appendix A at the rear of this manual. Also, see Subsection 3.2I.

SECTION II
INITIAL CHECKS

2.1 INTRODUCTION

The following procedure is provided to facilitate initial performance checking of the Model 124A. In general, the procedure should be performed after inspecting the instrument for shipping damage (any noted to be reported to the carrier and to Princeton Applied Research Corporation), but before using it experimentally. IN THE CASE OF UNITS HAVING A DIGITAL PANEL METER, IT IS IMPORTANT THAT THE PHRASE "meter full scale" BE PROPERLY INTERPRETED. READ THE PARAGRAPH BEGINNING WITH "In reading the display ---" ON PAGE III-24 BEFORE PROCEEDING WITH THE INITIAL CHECKS. Should any difficulty be encountered in carrying out these checks, contact the factory or one of its authorized representatives.

2.2 EQUIPMENT NEEDED

(1) General-purpose oscilloscope.

(2) Oscillator, having any 1 kHz repetitive waveshape that crosses its mean exactly twice each cycle and having a pk-pk voltage anywhere between 100 mV and 3 V.

(3) Assorted BNC cables.

2.3 PROCEDURE (for digital units, see page III-24 before proceeding)

NOTE: This procedure must be performed in sequence.

(1) Install a preamplifier if not already installed.

(2) Check the rear-panel 115/230 switch. Make sure the number showing in the window corresponds to the line voltage to be used.

(3) Turn the front-panel Power switch OFF.

(4) Plug the line cord into the rear panel and wall receptacles.

(5) Set the front panel controls as follows.

Meter: Check mechanical zero. Adjust if necessary.
Preamplifier Input: DIRECT (if applicable)
Sensitivity switch: 1 mV (if using a Model 118 Preamplifier, set the Sensitivity switch to 10 mV)
Signal Channel Mode: FLAT
Signal Frequency dials: 4.05
Signal Frequency range: X100
Q switch: 100
Time Constant: 300 ms
Zero Offset potentiometer: fully counterclockwise
10 X Full Scale switch: OFF (center position)
Preamplifier Mode: A
Reference Frequency dials: Red, NORMAL, Digits, 4.05
Reference Frequency range: X100
Reference Mode: INTERNAL
Reference Level: 10 (cal.)
Phase potentiometer: 90 (9 full turns)
Phase switch: 180°
Function switch: ACVM
Calibrator switch: 1 mV

(6) Connect a cable between the Calibrator BNC jack and the preamplifier's A input.

(7) Turn the Power on and wait five minutes for warmup. **NOTE:** In an actual measurement application, allow one hour warmup for optimum performance.

(8) The Meter should read to the right.

(9) Set the Signal Mode switch to BANDPASS.

(10) Adjust the right-most Signal Channel frequency dial for a peak on the meter (approximately full scale).

(11) Switch the Q switch back and forth between 100 and 10% ENBW positions, and adjust the Notch (front-panel screwdriver adjustment) for minimum change between the two positions (less than 1% of F.S. change). After adjusting, leave the Q at 100.

(12) Set the Sensitivity ADJ. potentiometer (front-panel screwdriver adjustment) for exactly full scale meter indication.

(13) Set the Signal Mode switch to NOTCH. The meter should now indicate 40% ±15% of positive full scale.

(14) Set the Signal Mode switch to LOW PASS. The meter should now indicate 90% ±10% of positive full scale.

(15) Set the Signal Mode switch to HIGH PASS. The meter should continue to indicate 90% ±10% of positive full scale.

(16) Set the Signal Mode switch to BANDPASS.

This completes the Signal Channel checks. The Phase and Function checks follow.

(17) Set the Function switch to LO DRIFT.

(18) Set the Q Selector to 10% ENBW. Then adjust the Phase potentiometer for zero on the meter. Lock the potentiometer.

(19) Set the Phase switch to 270°. Adjust, if needed, the Sensitivity screwdriver control for plus full-scale meter indication.

(20) Set the Phase switch to 90°. The meter should indicate minus full-scale ±2%.

(21) Set the Phase switch to 0°. The meter should indicate zero ±2% of full scale.

(22) Set the Phase switch back to 270°. (The meter should indicate positive full scale ±1%).

(23) Set the Function switch to NORMAL. The meter should remain at positive full scale (±1%).

(24) Set the Function switch to HI DYN RANGE. The meter should continue to indicate positive full scale (±1%).

This completes the Phase and Function checks. The Reference Channel checks follow.

(25) Connect the Signal Generator to the Reference Channel's In jack. The pk-pk voltage can be anywhere between 100 mV and 3 V. Set the signal generator's frequency to approximately 1 kHz.

(26) Monitor the Reference Channel's Out jack with the oscilloscope. The waveform should be a 28 V (±2 V) pk-pk sinewave.

(27) Set the Frequency Mode switch to EXT. The Ref. Unlock light should come on. Observe the frequency of the oscilloscope waveform. It should have begun increasing as soon as the Mode switch was set to EXT. After several seconds, it should stop increasing. When it stops, the Ref. Unlock light should go out. (The meter will go to Zero also.)

(28) Note the frequency of the oscilloscope waveform. Then place the Reference Mode switch to f/2; the frequency should double.

This completes the Reference Channel checks. The Sensitivity Range checks follow.

(29) Set the Reference Mode switch back to INT.

(30) Set both the Signal Channel Sensitivity and the Calibrator output to 100 mV. The meter should remain at full scale ±1%. **NOTE:** Calibrator to 10 mV with Model 118 Preamplifier.

(31) Progressively rotate the Sensitivity and Calibrator switches one position at a time, in a counterclockwise direction. The meter should remain at full scale ±2% if the two switches are in corresponding positions. If the meter wavers too much in the low nV settings, increase the Time Constant to 10 seconds. (If using the Model 118 Preamplifier, remember that the instrument is 10 X more sensitive than is marked on the Sensitivity switch, and set the switches accordingly. Just be sure to check all sensitivities for which Calibrator voltages are available.)

This completes the Sensitivity Range checks. The Output Offset and Overload checks follow.

(32) Remove the Input signal to the Preamplifier.

(33) Set the Time Constant switch to 300 ms.

(34) Set the Sensitivity switch to 1 mV.

(35) Place the 10 X Full Scale switch to "—".

(36) Adjust the Offset potentiometer for exactly one turn clockwise. The meter should indicate positive full scale ±2%.

(37) Place the 10 X Full Scale switch in "+". The meter should now indicate negative full scale ±2%.

(38) Increase the Offset potentiometer setting to 1.6 turns. The Overload light should come on.

(39) Return the 10 X Full Scale switch to its neutral off position. The Overload light will go out.

This completes the Initial Checks. If the instrument performed as indicated, one can be reasonably sure that it is operating properly.

SECTION III
OPERATING INSTRUCTIONS

3.1 BLOCK DIAGRAM DISCUSSION

3.1A INTRODUCTION

Before discussing the actual operation of the Model 124A, let us examine a functional block diagram to better understand what each adjustment does, and how the various adjustments relate to or influence one another. The functional block diagram is located on the following page. Schematics and a chassis wiring diagram are included at the back of this manual.

The Synchronous Detector is the heart of the instrument, around which are situated the Signal Channel, the Reference Channel, and the Output Amplifier. The Signal Channel amplifies and filters the signal, cleaning it up as much as possible before passing it along to the Synchronous Detector. The Reference Channel controls synchronization.

3.1B SIGNAL CHANNEL

Preamplifier

Four models of plug-in preamplifier are available: the Model 116, the Model 117, the Model 118, and the Model 119*. Together they cover the whole frequency spectrum from near dc to 210 kHz, each model having the best kind of input circuit for optimum low-noise performance in its frequency range. The Model 116 may be considered the general-purpose choice, performing well in most situations. Specific data for the four preamplifier models is given in the specifications in Section I and in the discussions in Subsection 3.2.

Selective Amplifier

The Selective Amplifier functions as a variable Q filter, which may be operated in the high-pass, low-pass, notch, bandpass, or flat mode. Because rms noise amplitude is a direct function of bandwidth, much of the noise can be rejected in this stage by filtering out all but the band containing the wanted signal. In addition, odd harmonics, to which the Synchronous Detector is sensitive, can be eliminated.

Intermediate Amplifier

The Intermediate Amplifier provides additional gain so that the signal-plus-noise applied to the Synchronous Detector is as large as possible without overload. Presenting a large signal to the Synchronous Detector minimizes, for a given overall sensitivity, the noise and dc drift contributed by the Output Channel.

*Also available are the Model 184 Photometric Preamplifier and the Model 185 Single Ended Low Noise Preamplifier. These two preamplifiers were developed later than the preamplifiers discussed in this manual. For information and specifications, contact the factory. Separate instruction manuals are provided for the Models 184 and 185.

The Intermediate Amplifier is ac coupled, thereby eliminating dc drift problems in the Signal Channel.

3.1C REFERENCE CHANNEL

Voltage-Controlled Oscillator

The VCO either locks onto a synchronizing signal from the experiment, or provides a synchronizing signal to the experiment. The VCO drives the Synchronous Detector so that the experiment and the Model 124A are properly synchronized.

The VCO automatically phase-locks to any kind of reference waveform having a frequency within approximately two decades of the Reference Channel band setting, the only requirements being that the waveform cross the mean twice (only) each cycle, that it have a pk-pk amplitude of at least 100 mV, and that it be synchronized with the signal of interest. The VCO can also lock to the second harmonic of the reference signal, if desired.

It is important to realize that, even though the VCO will phase lock over a two decade frequency range, the boundaries of the two decades are determined by the setting of the Oscillator Range switch. The tracking range corresponding to each position of the switch is as follows.

X1	0.2 Hz to 21 Hz
X10	2 Hz to 210 Hz
X100	20 Hz to 2.1 kHz
X1K	200 Hz to 21 kHz
X10K	2 kHz to 210 kHz

In the Internal mode, a sinewave output from the Oscillator is provided for synchronizing the experiment. The Oscillator can free run accurately at any selected frequency from 200 mHz to 210 kHz. Also, the Oscillator's frequency may be controlled by a voltage applied to a rear-panel jack.

Phase Controls

The Phase Controls combine quadrature outputs from the VCO such that the resultant sinewave presented to the Synchronous Detector has the desired phase relation to the reference (sync input and/or output).

3.1D SYNCHRONOUS DETECTOR

The Synchronous Detector inverts the polarity of the part of the input signal corresponding to the negative excursion of the phase-shifted sinewave from the VCO, and passes the remaining waveform uninverted. Therefore, the dc component of the resultant waveform is proportional to the value of signal at the same frequency and phase as the phase-shifted VCO sinewave. Because the detected signal still has noise on it, RC low-pass filters that follow the synchronous detector are used to eliminate all but the dc component representing the wanted signal. A dc amplifier

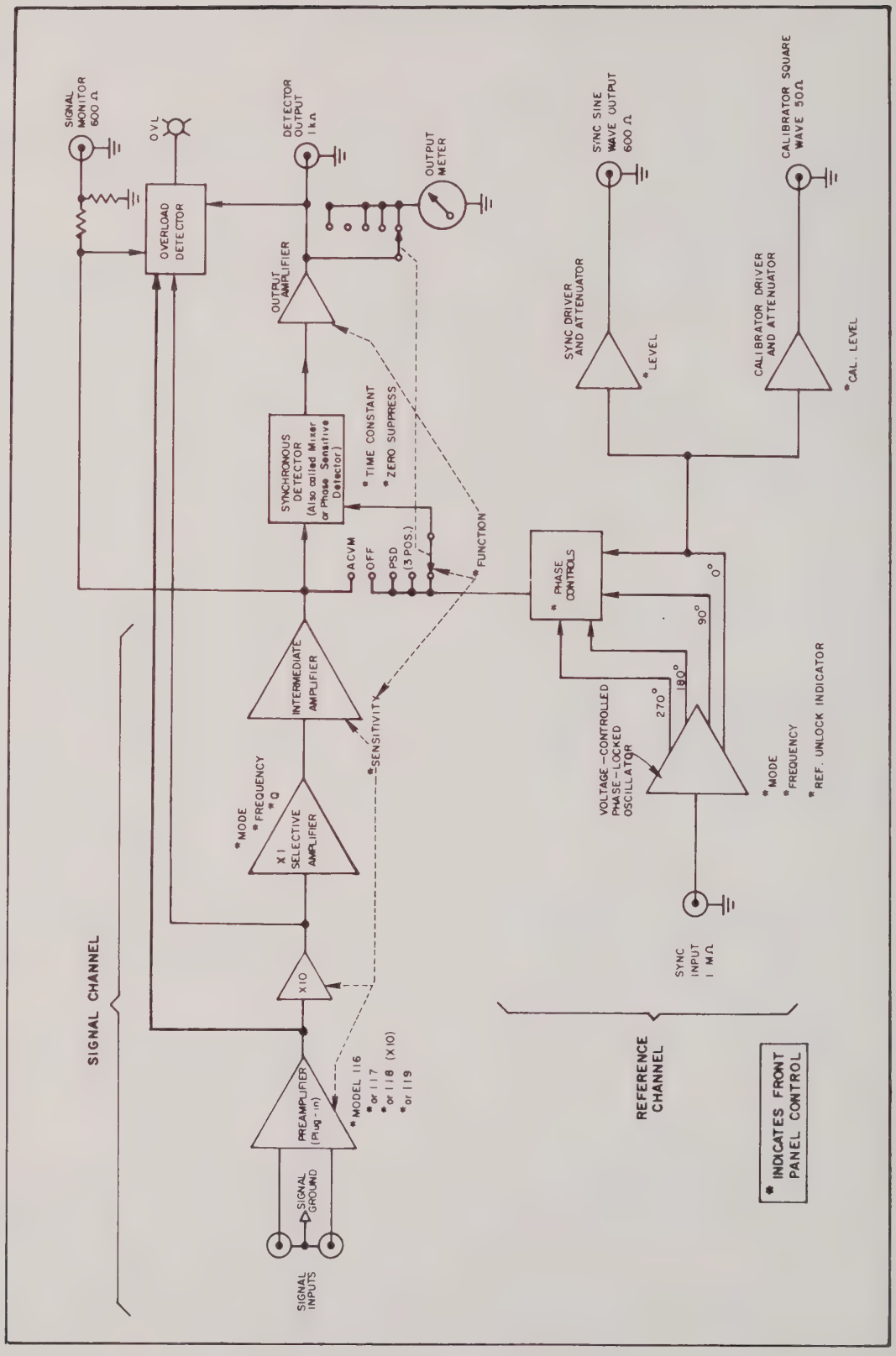


Figure III-1. FUNCTIONAL BLOCK DIAGRAM, MODEL 124A

following the Detector provides the final gain. This amplifier drives the output connector and panel meter.

To determine the amplitude of a signal, the operator simply adjusts the phase controls for maximum meter indication. The meter directly indicates the signal amplitude. In addition, after making this adjustment, the phase of the input signal relative to the reference may be accurately read from the Phase dial.

3.2 SIGNAL CHANNEL OPERATION

3.2A INTRODUCTION

The function of each control is indicated in Figure I-1. Instead of repeating the information given there, this subsection provides additional background information for setting the controls to their optimum positions.

To set up the Signal Channel controls, the operator must know the frequency or frequency range of the signal. It also helps if he knows the expected amplitude, the amount and type of noise obscuring the signal, and the signal-source impedance.

The stimulus to the experiment is often chopped. Detecting at this chopping frequency eliminates all but the signal related to the stimulus. Princeton Applied Research Corporation manufactures several models of light chopper which are ideally suited to photo-detection applications. Chopping the source light finds its analog for other kinds of experiments in chopping the stimulus dc, rf, ac, sound, heat, etc. The chopping frequency range must be fairly well known, and sync signals must be available. Of course, the frequency of interest may not be a result of chopping at all. The chopping technique discussed here merely serves as an example.

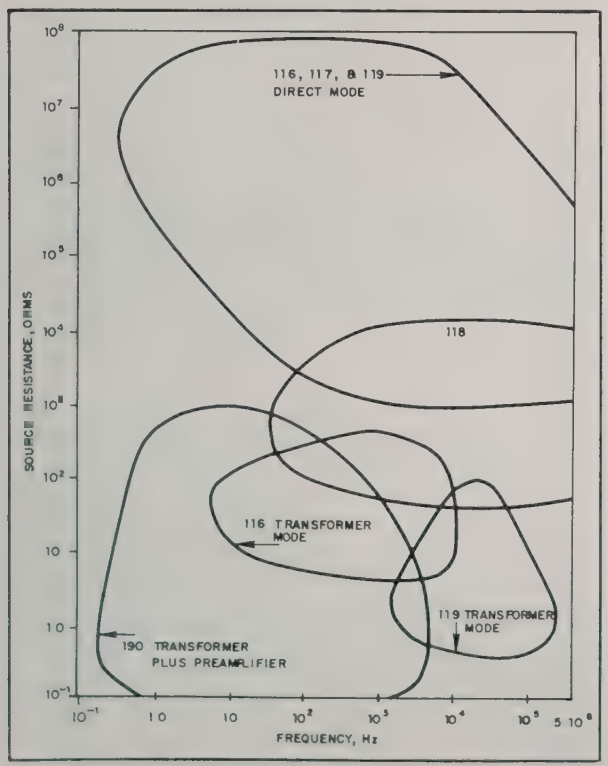


Figure III-2. OPTIMUM PERFORMANCE REGIONS OF THE PREAMPLIFIERS

3.2B PREAMPLIFIER CHOICE

Four models of plug-in preamplifiers are available, each one providing optimum low-noise performance over a given input-resistance v. frequency range, as shown graphically in Figure III-2. Two of these preamplifiers can be operated both in direct and transformer-coupled modes, and all four of them can be operated single-ended or differentially. A Model 190 transformer can be used with any of the plug-ins to improve low-frequency performance when working from low source impedances. Use of the Noise Figure contours is discussed in Subsection 3.2G.

Another consideration in selecting a preamplifier is its ability to amplify without distortion. If an experiment requires measurement of low level harmonics in the presence of a high level fundamental, it is important that the preamplifier not add significant harmonic signals by non-linearly amplifying the fundamental. Except in the case of the Model 118 Preamplifier, the distortion generated by preamplifiers operating in the direct mode is so small as to be unmeasurable using conventional methods. The Model 118 can, however, distort more under certain conditions, as indicated in Figure III-3. In the transformer mode, both phase shift and distortion must be measured for each individual operating condition.

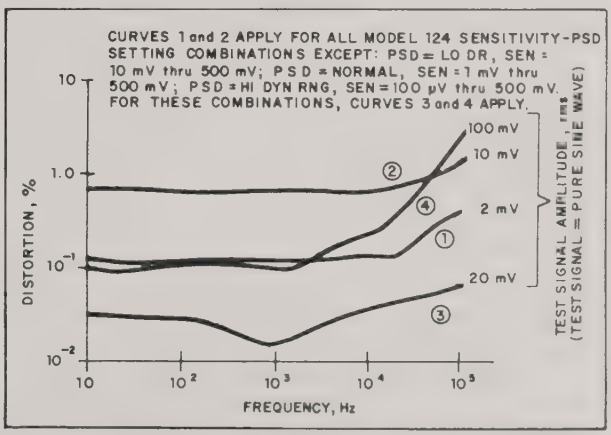

CURVES 1 and 2 APPLY FOR ALL MODEL 124 SENSITIVITY-PSD SETTING COMBINATIONS EXCEPT: PSD = LO DR, SEN = 10 mV thru 500 mV; PSD = NORMAL, SEN = 1 mV thru 500 mV; PSD = HI DYN RNG, SEN = 100 µV thru 500 mV. FOR THESE COMBINATIONS, CURVES 3 and 4 APPLY.
100 mV
10 mV
2 mV
20 mV
TEST SIGNAL AMPLITUDE, rms (TEST SIGNAL = PURE SINE WAVE)
DISTORTION, %
FREQUENCY, Hz

Figure III-3. DISTORTION v. FREQUENCY, MODEL 118

3.2C GROUNDING

In any system processing low-level signals, proper grounding to minimize the effects of ground-loop currents, usually at the power frequency, is an important consideration. With the exception of the Models 184 and 185, all of the Model 124A preamplifiers allow both differential and single-ended operation. Two properties of these preamplifiers allow them to achieve a high degree of immunity to ground-loop currents. In differential operation, their extremely high common-mode rejection assures an almost total rejection of

unwanted signals which appear at both inputs at the same phase and amplitude, the usual case for ground-loop interference. In single-ended operation, their unique input grounding system, in which signal ground "floats" off chassis ground by ten ohms, assures a high degree of rejection. As a result of this ten ohm ground, ground loop signals are effectively attenuated by the ratio of ten ohms to the braid resistance (typically 20 mΩ) of the cable carrying the signal from the source to the input.

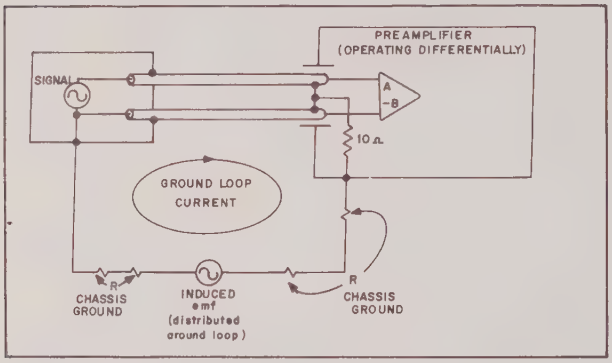


Figure III-4. GROUND-LOOP VOLTAGE REJECTION USING DIFFERENTIAL INPUTS

Despite the immunity granted by the ten-ohm ground, far better rejection can be achieved in processing a signal from a single-ended source if one operates differentially as shown in Figure III-4. In single-ended operation, the preamplifier "sees" the potential difference between the center conductor of the cable and the braid, and any ground loop signal on the braid can be attenuated but not rejected altogether. In differential operation, the preamplifier "sees" the *potential difference* between the "A" Input and the "B" Input. Ground-loop signal current flowing in the braid is of no consequence.

However, when operating differentially, it is important to assure that common-mode interference arising in ground loops is just that, i.e., without a significant differential component. This should not prove a problem as long as both signal cables are the same length and follow the same path.

3.2D REMOTE PREAMPLIFIER ADAPTER

In situations where very low signal levels are encountered, it may be desirable to operate the preamplifier very close to the signal source to reduce noise and stray pickup, while leaving the main Lock-In unit at a convenient operating location. This may be accomplished with the accessory Remote Preamp Adapter, Model 183 (see Figure III-5).

3.2E SINGLE-ENDED, DIFFERENTIAL, AND TRANSFORMER INPUTS

All of the preamplifiers have switch-selectable single-ended/differential inputs. The differential inputs can be used to combine signals (A – B) as well as to provide common mode rejection.

Impedance matching of the source and loading considerations often require that the input impedance of the preamplifier be known. However, do not confuse impedance matching with optimum input impedance for low-noise operation. The latter is discussed in detail in Subsection 3.2G.

The input impedance of each preamplifier (direct mode) is: Model 116, 100 megohms 20 pF; Model 117, 100 megohms 20 pF; Model 118, 10 kilohms 170 pF; Model 119, 100 megohms 20 pF. These impedances are for each input to ground. In the differential mode, for all models except the 118, the impedance from input to input is twice that stated (i.e., R double, C half). For the Model 118, the input impedance for the differential mode is about the same as the impedance for single-ended operation. In addition, the diff. input impedance of the Model 118 varies with sensitivity and the setting of the PSD switch (common-mode input impedance is 25 kΩ for all combinations). In the Lo Drift mode, on the 10 mV through 500 mV sensitivity ranges, the input impedance is about three times that specified. In the Normal mode, the higher impedance applies to the 1 mV through 500 mV ranges, and in the Hi Dynamic Range mode, it applies to the 100 μV through 500 mV ranges.

For the preamplifiers having internal transformers, the Input switch operates the same for the transformer as it does for direct. The input impedance is very low in the transformer mode.

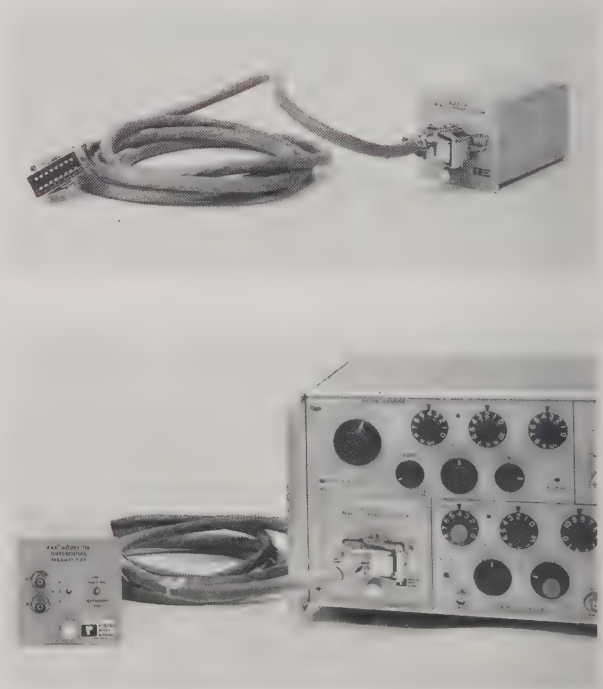


Figure III-5. MODEL 183 REMOTE PREAMPLIFIER ADAPTER

If an external transformer, such as the Model 190, is used, it is connected to the preamplifier single ended. The transformer alone provides sufficient common-mode rejection (Figure III-6).

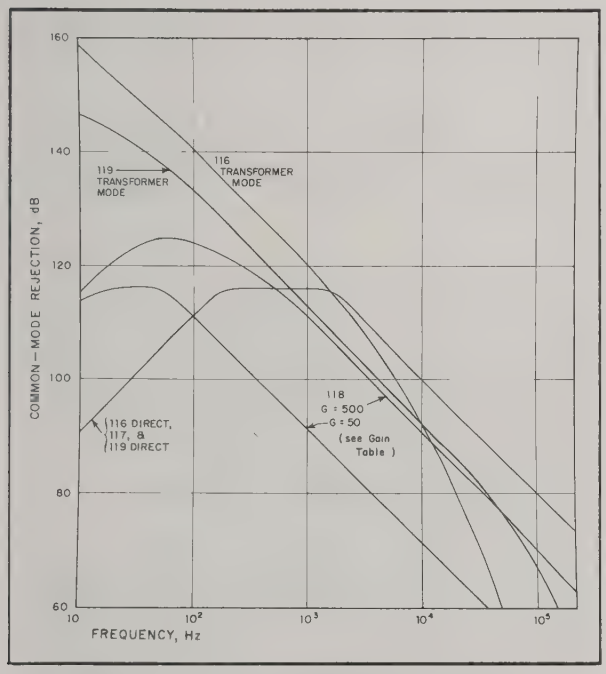


Figure III-6. TYPICAL COMMON-MODE REJECTION

3.2F COMMON-MODE REJECTION

Figure III-6 illustrates the common-mode rejection characteristics of the preamplifiers in both (differential) direct and transformer mode operation. Note that the CMR is much higher for the transformer mode than for the direct mode.

3.2G NOISE AND SOURCE RESISTANCE

Best preamplifier performance is realized under those conditions where the overall signal-to-noise ratio is least degraded. In many instances, the thermal noise generated by the signal source resistance is the dominant factor in determining the input signal-to-noise ratio. In this respect, amplifier noise performance can be specified by the amount of noise the amplifier adds to the amplified source thermal noise; expressed in decibels this is called the "Noise Figure":

(1) Noise Figure =

$$20 \log_{10} \frac{\text{total rms noise voltage at the amplifier output}}{\text{gain x source thermal noise voltage (rms)}} \text{ dB}$$

where the Source Thermal Noise = $\sqrt{4KTR\Delta f}$ volts rms = E, and

K = Boltzmann's constant, 1.38×10^{-23} joules/K
T = absolute temperature in kelvins
Δf = equivalent noise bandwidth in Hz
R = source resistance in ohms

The total output noise may be converted to an equivalent input noise by dividing by the amplifier gain. The noise figure, expressed in these terms, becomes:

(2) Noise Figure =

$$20 \log_{10} \frac{\text{total rms noise voltage referred to amp. input}}{\text{source thermal noise voltage (rms)}} \text{ dB}$$

Each amplifier has its own characteristic noise figure, which varies as a function of frequency and source resistance. These figures are obtained experimentally and plotted graphically. Figures III-7 through III-11 include typical sets of noise figure contours for the Models 116, 117, 118, and 119 operating in direct and transformer modes.

Using the applicable set of contours, the total equivalent rms input noise can be determined:

(3) Total equivalent rms input noise voltage = source thermal noise x antilog NF/20 volts rms.

Notice from these equations that the bandwidth must be specified; usually determined by the external circuitry and/or the amplifier bandwidth. Figures III-7 through III-11 include response curves of the preamplifiers from which the bandwidth may be obtained (in the direct modes, the bandwidth is wider than the widest Tuned Amplifier Bandwidth). However, a more interesting place to determine noise level is at the output of the Tuned Amplifier. Because there is gain ahead of the Tuned Amplifier, its noise contribution referred to the input is negligible compared to that of the preamplifier. In computing the total equivalent input noise of the signal channel, the operator only need use a Δf as determined by the Selective Amplifier setting (with preamplifier limitations considered).

Although the discussion of noise considerations is not complete, a simple example at this point illustrates the use of the foregoing equations and illustrates how a transformer can, on occasion, improve the signal-to-noise ratio:

Suppose that one intended to operate the Model 124A in an experiment having a source impedance of 10 ohms. Further suppose the signal frequency to be 5 kHz. A Model 116 Preamplifier is chosen. In order to see how a transformer can improve low-noise performance, the noise for the Direct mode is first calculated. Then the noise for transformer operation is calculated and the results compared. Since the source thermal noise contributing to the total noise is dependent on bandwidth, the Signal Channel bandwidth is minimized by setting the Mode switch in the Bandpass position and the Q switch to the 10% ENBW (equivalent noise bandwidth) position. The Frequency controls are set to 5 kHz. In addition, other control settings are made according to instructions in other parts of this section. The source thermal noise in this case is:

$$E = \sqrt{4KTBR}$$

where:

k is Boltzmann's constant = 1.38×10^{-23} joules/K

T is the absolute temperature of the source in kelvins, presumed to be 290 K for the example

B is the noise bandwidth, = 500 Hz with Q switch set to 10% ENBW and Frequency Controls set to 5 kHz

R is the source resistance in ohms, given as 10 ohms.

Thus:

$$E = \sqrt{4 \times 1.38 \times 10^{-23} \times 2.9 \times 10^{2} \times 10 \times 5 \times 10^{2}}$$

$$= 8.9 \times 10^{-9} \text{ V rms}$$

From Figure III-7, the noise figure for the Model 116 at a center frequency of 5 kHz and a source impedance of 10 ohms is 20 dB. Substituting E and this NF into equation (3), we get:

total equivalent rms input noise

$$= 8.9 \times 10^{-9} \times 10^{20/20}$$

$$= 89 \text{ nV rms}$$

With a transformer inserted between the signal source and the amplifier input we can increase the effective source impedance to a value that reduces the noise figure to less than 0.1 dB. From Figure III-7, the source resistance should be about 200 kilohms. The transformer turns ratio required for this impedance increase is $\sqrt{R_2/R_1} = 141$. The thermal source noise at the amplifier input is equal to the noise generated by the 10 ohm source multiplied by the turns ratio. With a noise figure near zero, the total equivalent rms input noise is also equal to the noise generated by the 10 ohm source multiplied by the turns ratio, = 1.25 μV rms. Although the numerical value of equivalent input noise is much larger than before, the signal-to-noise ratio is substantially increased. This can be seen by considering all of the transformed source signal voltage as appearing at the amplifier input terminals, possible because the Model 116's input resistance is much larger than 200 kilohms presented by the transformer; the signal-to-noise ratio is equal to the maximum possible value: $e_{sig}/(E \times \text{antilog NF}/20) \cong e_{sig}/E$, the noise contributed by the amplifier being negligible under the "near zero" noise figure conditions. In this example the transformer increases the signal-to-noise ratio by a factor of 10.

Signal-to-Noise Improvement Ratio

$$= 10^{(\text{NF unmatched} - \text{NF matched})/20}$$

$$= 10^{(20 - .1)/20} \cong 10^{1} = 10$$

However, because of the noise contributed by the transformer, and because a transformer influences bandwidth, the results obtained using a real transformer are never as good as the ideal theoretical results predicted in the example. Also, it is seldom convenient to obtain a transformer having exactly the ideal turns ratio. It is best, therefore, to use noise figure contours and amplitude transfer curves obtained empirically for the individual transformer. Figure III-8A is a set of NF contours for the Model 116's built-in transformer, and Figure III-8B is its amplitude transfer curves. The 10-ohm amplitude transfer curve indicates that in this example the transformer does not change the 10% equivalent noise bandwidth as set with the Q switch, so that E remains 8.9×10^{-9} V rms. The noise figure for a 10 ohm source at a center frequency of 5 kHz is about 1.5 dB, a vast improvement over the 20 dB NF obtained without the transformer.

NOTE: For optimum noise performance with a transformer, it is important that the transformer not be magnetized. See TRANSFORMER MAGNETIZATION AND DEGAUSSING, page III-11.

3.2H SELECTIVE AMPLIFIER

Noise other than source thermal noise is usually not wideband, and is often difficult to compute. Some kinds of noise can be dealt with very effectively using the Selective Amplifier. Examples include flicker (or 1/f) noise, non-synchronous signals arising from the experiment, non-synchronous signals from external pickup such as from the ac power line, fast transients, and harmonics of the reference frequency.

Reducing the noise level ahead of the mixer reduces the dynamic range demands on the mixer, thereby allowing signals to be measured which could not be measured otherwise.

The Selective Amplifier can be operated in five different modes: Flat, Bandpass, Notch, Low Pass, and High Pass. Figures III-14 through III-17 illustrate typical transfer characteristics of the Selective Amplifier for the last four of these modes. Ultimate attenuation of the four frequency-dependent curves exceed 80 dB.

In selecting filtering parameters, the operator must be careful to keep the signal frequency well within the passband selected, or to make the passband such as to accommodate the signal over the range that it will occupy. If phase is important, actual measurements of phase error over the frequency range of questionable phase accuracy would be best. These regions of questionable accuracy can be determined from the individual transfer characteristics of the preamplifier, transformer, and Selective Amplifier settings. The best thing to do is to keep the bandwidth wide enough so that phase and amplitude errors are not a problem. Phase control errors are discussed in Subsection 3.3B. Special procedures for making accurate phase measurements are given in Subsection 3.9. In particular, before operating in LOW PASS or HIGH PASS, the operator is advised to check Figures III-16 and III-17 to determine the amplitude responses in these modes as a function of Q.

"Equivalent noise bandwidth" is a concept applied to wideband noise. Although most bothersome noise is not wideband, but rather, coherent non-synchronous signal, equivalent noise bandwidth considerations are useful in helping to choose operating parameters. The concept of equivalent noise bandwidth arises from the fact that Noise Bandwidth is an unattenuated rectangular bandwidth, while

(continued on page III-13)

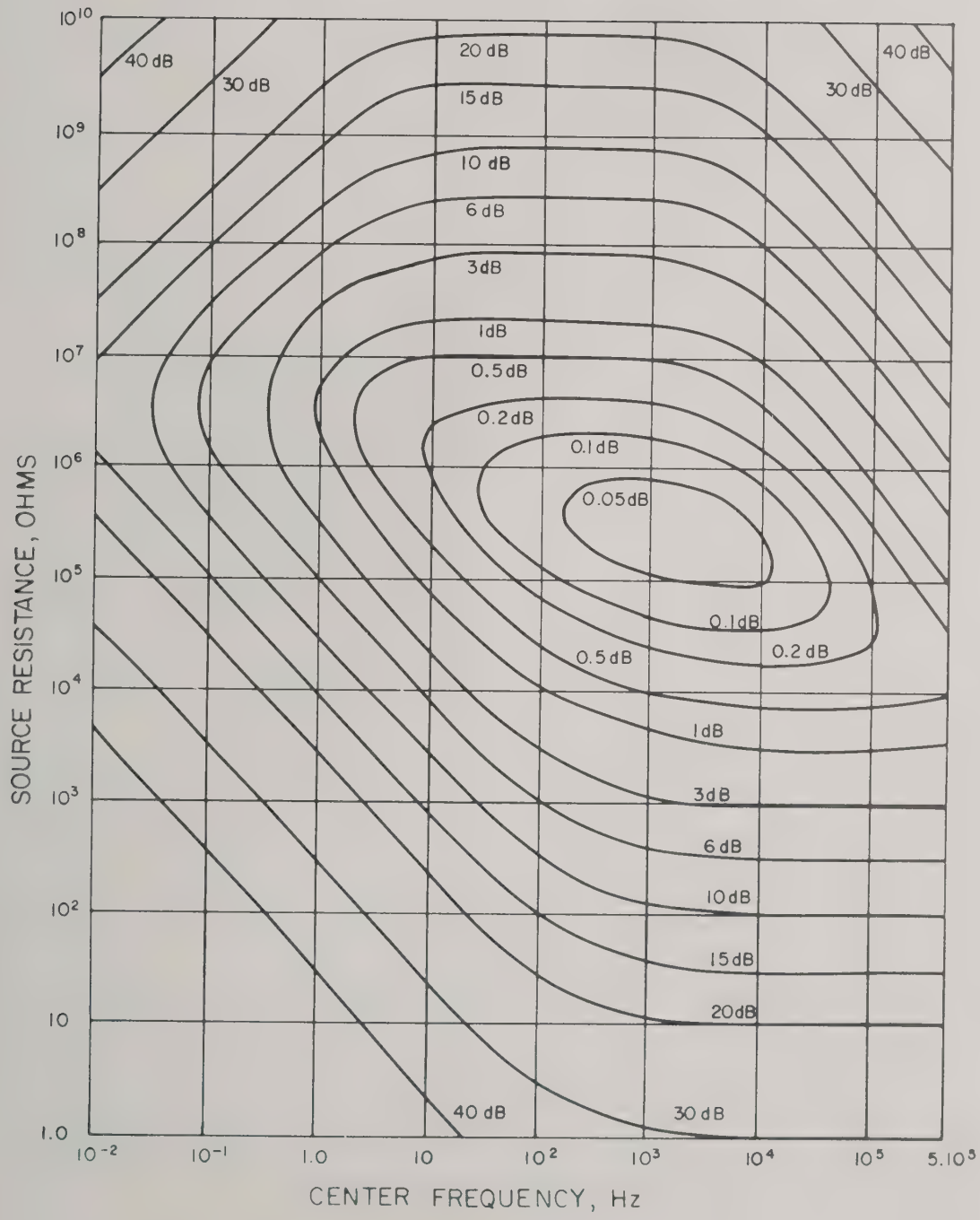


Figure III-7. TYPICAL NOISE FIGURE CONTOURS FOR MODEL 117, MODEL 116 (DIRECT) AND MODEL 119 (DIRECT)

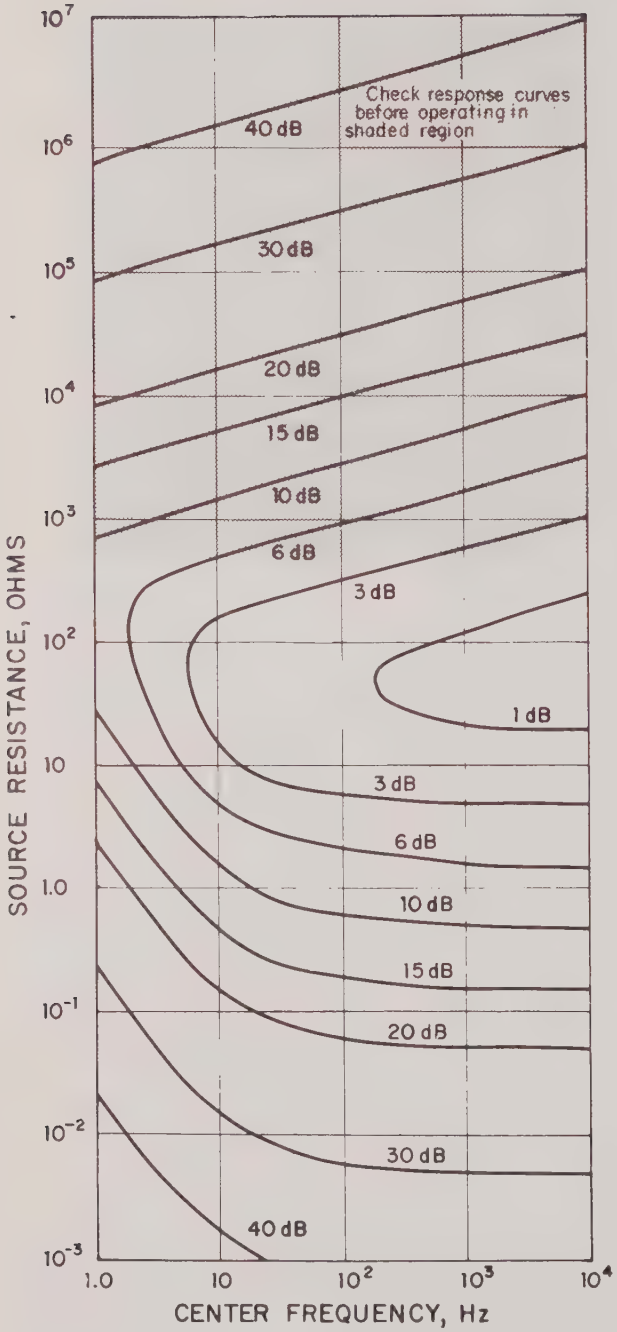


Figure III-8A. TYPICAL NOISE FIGURE CONTOURS FOR MODEL 116 OPERATING IN THE TRANSFORMER MODE

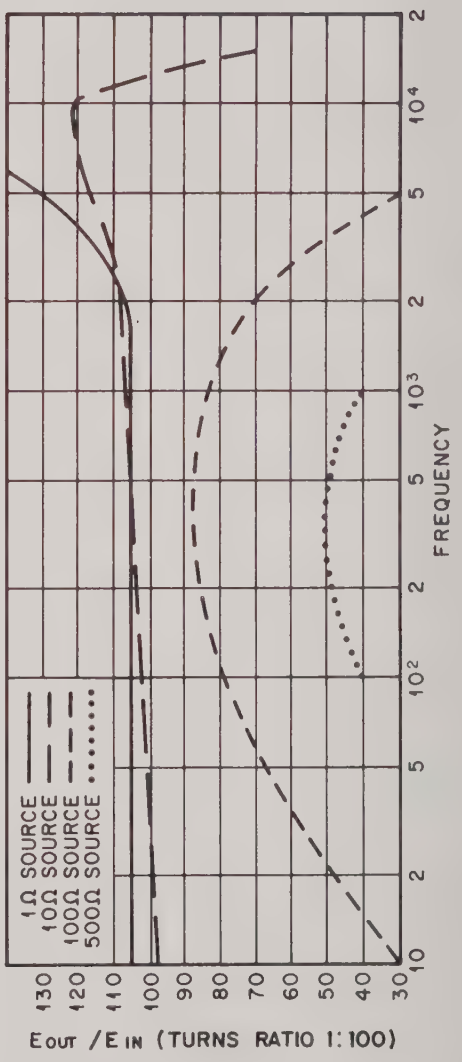


Figure III-8B. TYPICAL AMPLITUDE TRANSFER CURVES FOR MODEL 116 OPERATING IN THE TRANSFORMER MODE

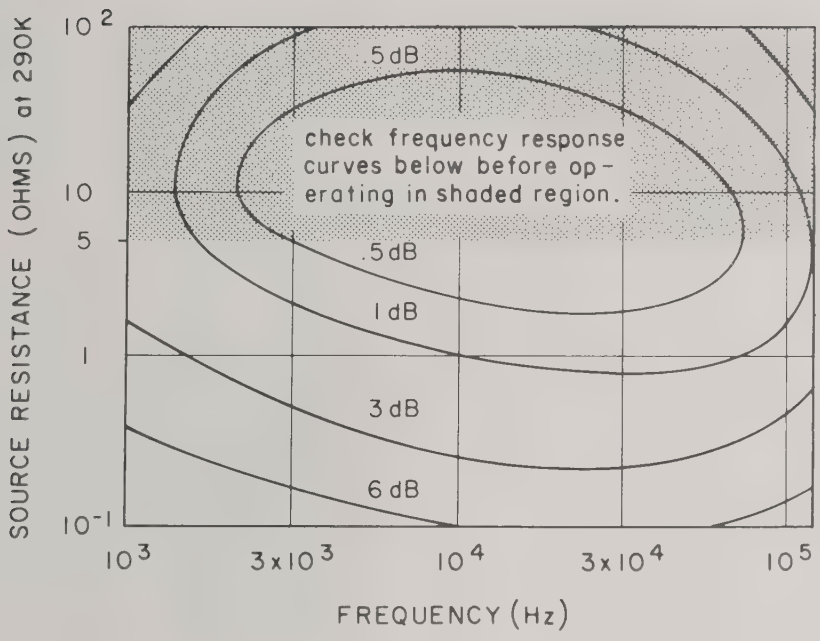


Figure III-9A. TYPICAL NOISE FIGURE CONTOURS FOR MODEL 119 OPERATING IN THE TRANSFORMER MODE

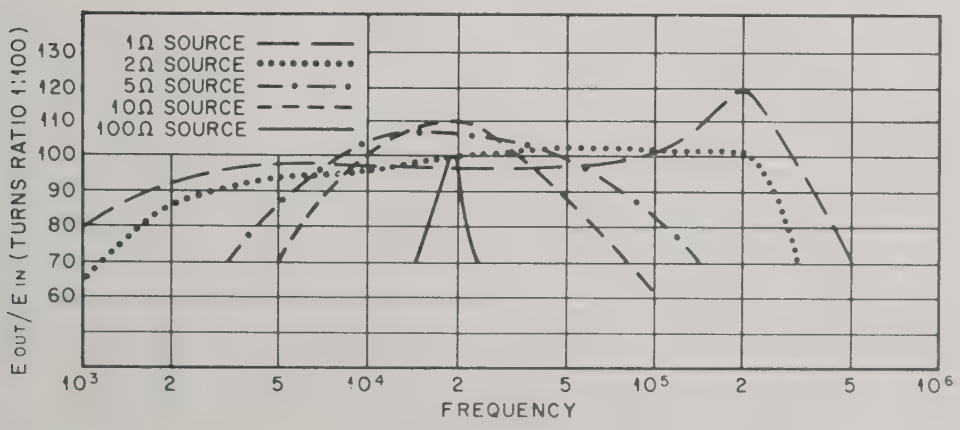


Figure III-9B. TYPICAL AMPLITUDE TRANSFER CURVES FOR MODEL 119 OPERATING IN THE TRANSFORMER MODE

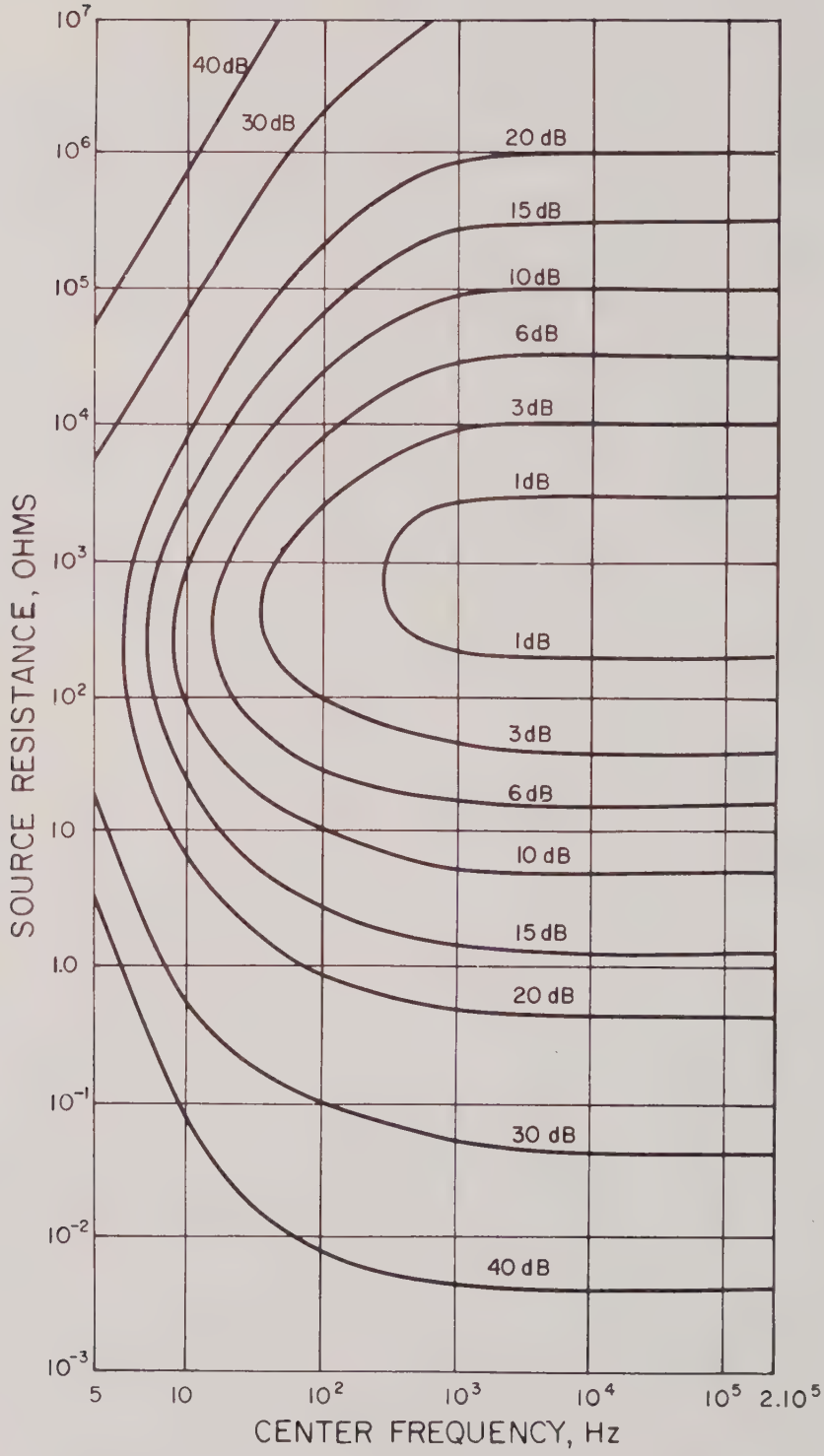


Figure III-10. TYPICAL NOISE FIGURE CONTOURS FOR MODEL 118

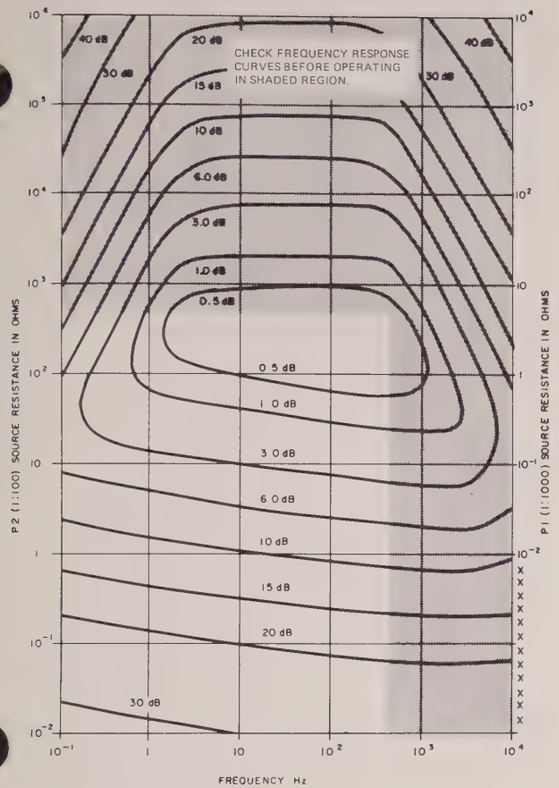


Figure III-11A. TYPICAL NOISE FIGURE CONTOURS FOR MODEL 190 TRANSFORMER-PLUS-PREAMPLIFIER

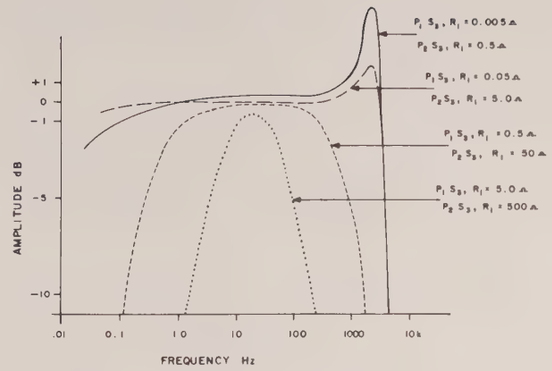


Figure III-11C. TYPICAL AMPLITUDE TRANSFER CURVES FOR MODEL 190

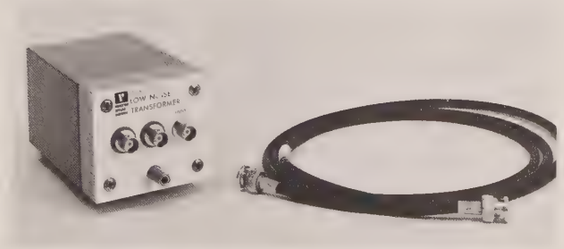


Figure III-11D. PHOTO OF MODEL 190 TRANSFORMER

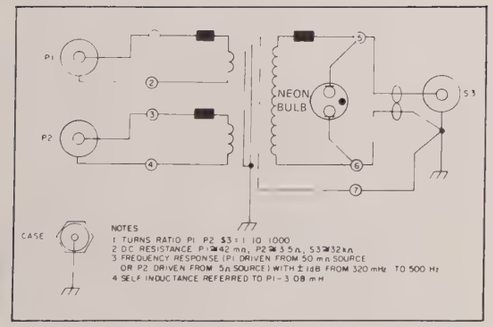

P1
P2
NEON BULB
S3
CASE
NOTES
1 TURNS RATIO P1-P2-S3 = 1-10-1000
2 DC RESISTANCE P1 ≅ 42 mΩ, P2 ≅ 3.5 Ω, S3 ≅ 32 kΩ
3 FREQUENCY RESPONSE (P1 DRIVEN FROM 50 mΩ SOURCE OR P2 DRIVEN FROM 5Ω SOURCE) WITH ± 1dB FROM 320 mHz TO 500 Hz
4 SELF INDUCTANCE REFERRED TO P1 = 3.08 mH

Figure III-11B. MODEL 190 WIRING DIAGRAM

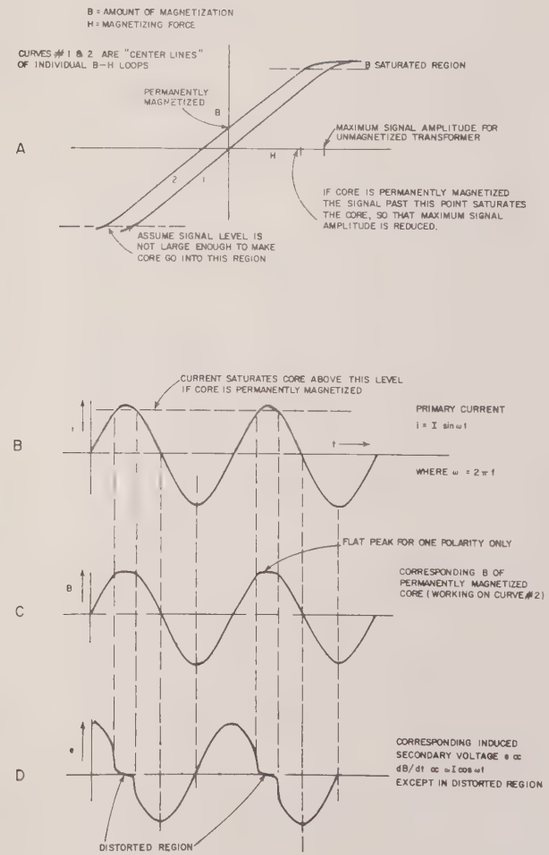

B = AMOUNT OF MAGNETIZATION
H = MAGNETIZING FORCE
CURVES #1 & 2 ARE "CENTER LINES" OF INDIVIDUAL B-H LOOPS
B SATURATED REGION
PERMANENTLY MAGNETIZED
B
A
H
MAXIMUM SIGNAL AMPLITUDE FOR UNMAGNETIZED TRANSFORMER
IF CORE IS PERMANENTLY MAGNETIZED THE SIGNAL PAST THIS POINT SATURATES THE CORE, SO THAT MAXIMUM SIGNAL AMPLITUDE IS REDUCED.
ASSUME SIGNAL LEVEL IS NOT LARGE ENOUGH TO MAKE CORE GO INTO THIS REGION
B
CURRENT SATURATES CORE ABOVE THIS LEVEL IF CORE IS PERMANENTLY MAGNETIZED
PRIMARY CURRENT
I = I sin ωt
WHERE ω = 2πf
FLAT PEAK FOR ONE POLARITY ONLY
C
CORRESPONDING B OF PERMANENTLY MAGNETIZED CORE (WORKING ON CURVE #2)
D
CORRESPONDING INDUCED SECONDARY VOLTAGE e ∝ dB/dt ∝ ωI cos ωt
EXCEPT IN DISTORTED REGION
DISTORTED REGION

Figure III-12. B-H CURVES AND WAVEFORMS

TRANSFORMER MAGNETIZATION AND DEGAUSSING

A magnetized transformer is usually noisy and microphonic compared to when it is not magnetized. This noise is the most objectionable characteristic of a magnetized transformer because it affects low level signals. The distortion of high level signals described here is not usually an operating problem because the signal is seldom so large. But high level distortion is useful in determining the condition of the core and in monitoring the degaussing process.

Avoid passing excessive dc current through the primary of these transformers. The transformer used in the Model 116 and the Model 190 transformer will permanently magnetize, and once magnetized, they will remain noisy until degaussed. More than 200 μA will magnetize the cores of these transformers. Do not measure winding resistance with an ohmmeter; ohmmeter current is sufficient to magnetize the core. The transformer used in the Model 119 has a ferrite core that will not permanently magnetize.

Notice from Figure III-12 that as long as the core is not magnetized the transformer works on a loop around center line #1 and the signal can be of maximum amplitude. If the core becomes permanently magnetized, the transformer works on a loop around center-line #2 (or equivalent for opposite polarity), and, if the same maximum signal current is applied, the core becomes saturated at one of the peaks of the signal. B, C, and D of Figure III-12 show a primary signal current, corresponding core magnetization, and corresponding induced secondary voltage for a magnetized-core transformer with large signal input. Notice that the permanently magnetized core saturates one polarity of signal and later another polarity.

To demagnetize (degauss) the transformer it is necessary to pass a relatively large current through the primary and slowly decrease this current. The secondary should be open (for the Model 116 disconnect pins 4 and 5 on the circuit board). The degaussing current is initially large enough to magnetize the core in the direction of each corresponding polarity excursion of this current, and overrides the initial magnetization. By decreasing the current each successive magnetization is slightly less than the preceding one of the same polarity. As the current is decreased the core is progressively demagnetized until it is no longer magnetized. Figure III-13A, B and C illustrate the waveforms as the transformer is demagnetized. IMPORTANT: Use a low FREQUENCY degaussing current, and do so, to avoid inducing excessive voltage in the secondary. Monitor the secondary with a high impedance scope. The Model 190's transformer can be protected with a neon lamp, but the Model 116's transformer could suffer insulation breakdown if the operator is not careful.

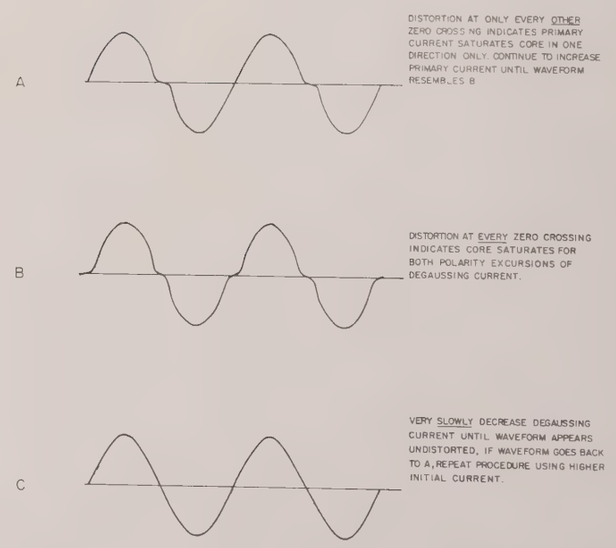
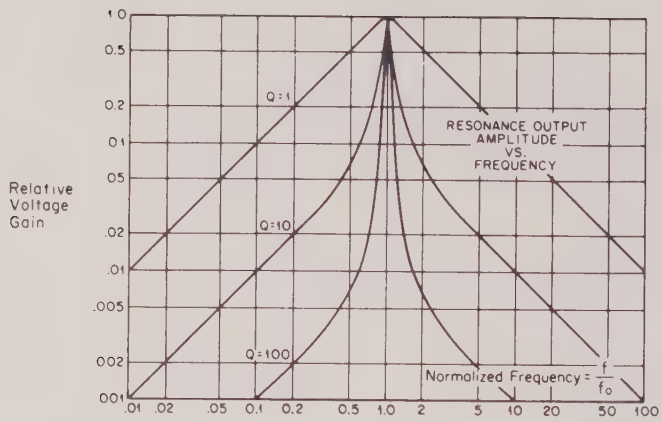

A
DISTORTION AT ONLY EVERY OTHER ZERO CROSSING INDICATES PRIMARY CURRENT SATURATES CORE IN ONE DIRECTION ONLY. CONTINUE TO INCREASE PRIMARY CURRENT UNTIL WAVEFORM RESEMBLES B
B
DISTORTION AT EVERY ZERO CROSSING INDICATES CORE SATURATES FOR BOTH POLARITY EXCURSIONS OF DEGAUSSING CURRENT.
C
VERY SLOWLY DECREASE DEGAUSSING CURRENT UNTIL WAVEFORM APPEARS UNDISTORTED, IF WAVEFORM GOES BACK TO A, REPEAT PROCEDURE USING HIGHER INITIAL CURRENT.

Figure III-13. DEGAUSSING WAVEFORMS



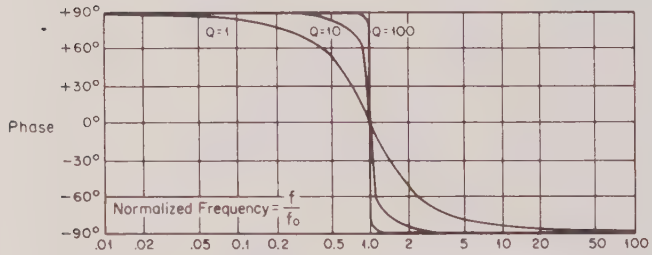
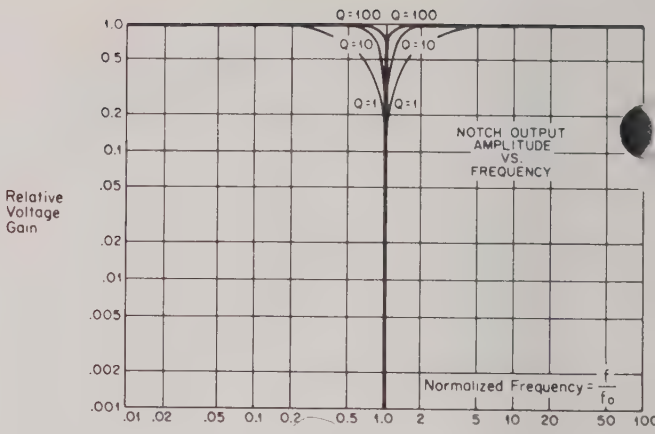


Figure III-14. MODEL 124A BANDPASS CHARACTERISTICS



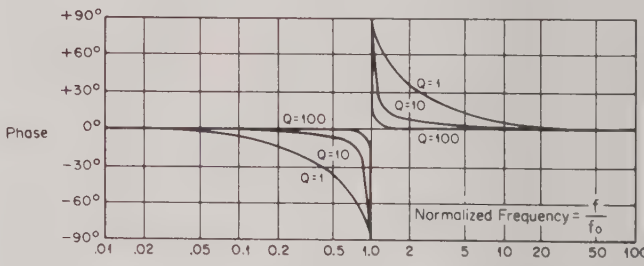
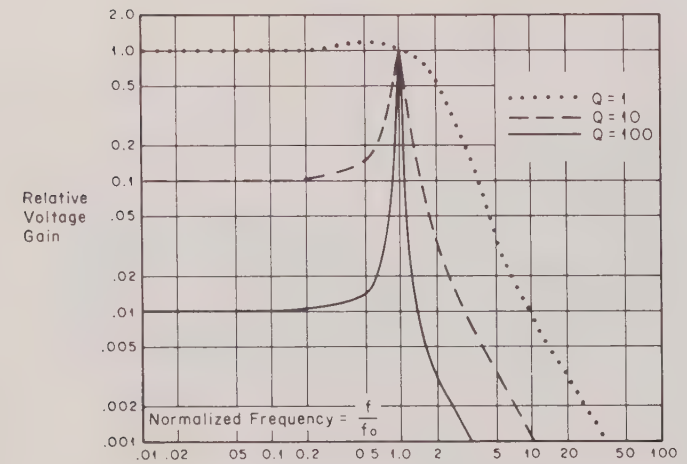


Figure III-15. MODEL 124A NOTCH CHARACTERISTICS



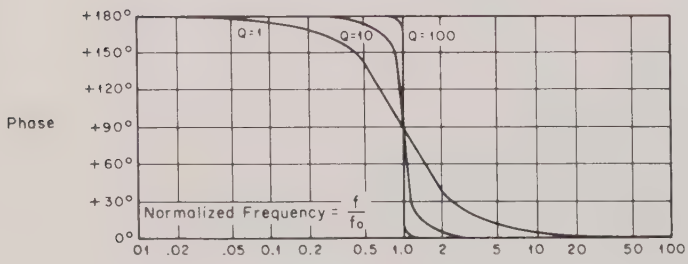
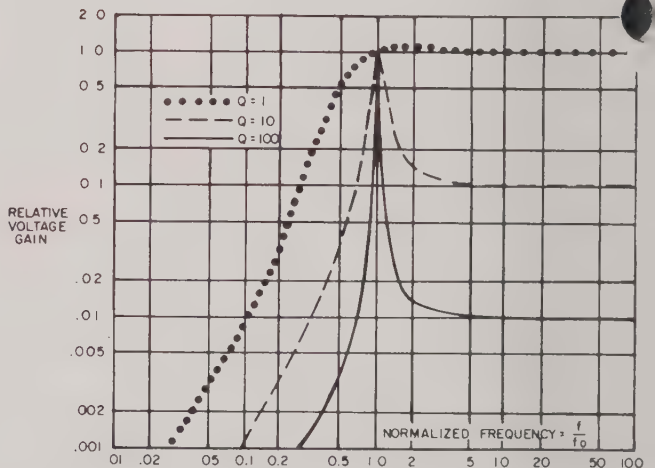


Figure III-16. MODEL 124A LOW PASS CHARACTERISTICS



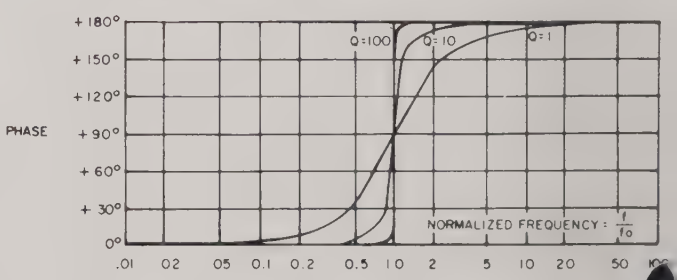


Figure III-17. MODEL 124A HIGH PASS CHARACTERISTIC

signal bandwidths are specified as the number of Hz between two points of given attenuation on the response characteristic. Suppose one had a filter with a given signal bandwidth, and some amount of wideband noise, measured in volts/$Hz^{1/2}$, were applied to it. At the output of the filter one would measure some amount of rms noise voltage. The equivalent noise bandwidth of the filter is defined as the unattenuated bandwidth of an equivalent theoretical (but physically impossible) perfectly sharp filter that, with the same wideband input noise, yields the same rms output noise amplitude as the filter of interest.

On the Model 124A, notice from Figures III-16 and III-17 that the frequency dial setting corresponds to the last point for which the response is unity. However, signal bandwidth is ordinarily taken between 3 dB (half-power) points on the rolloff characteristic. For the Bandpass mode, 3 dB bandwidth is obtained from the standard formula for Q ($Q = f_o/\Delta f_{3dB}$).

Calculating ENBW (equivalent noise bandwidth) from the 3 dB bandwidth is frequently complicated. However, an exact conversion is seldom required. The 3 dB bandwidth and ENBW are of the same order of magnitude with the generally applicable (and sufficiently accurate) conversion factor being $\pi/2$. The FLAT mode ENBW is simply the 3 dB upper limit multiplied by $\pi/2$. The same applies to a close approximation for operation in the NOTCH mode. In LOW PASS operation, the ENBW is the product of the 3 dB down frequency as determined from Figure III-16 times $\pi/2$. In HI PASS operation, the ENBW is $\pi/2$ times the square of the upper limit divided by the sum of the upper limit frequency and the 3 dB down frequency as determined from Figure III-17. In BANDPASS operation, the ENBW is $\pi/2$ times f/Q, where f is the selected tuned frequency and Q is that selected by the Q switch. For all modes, bear in mind that if the signal bandwidth is limited ahead of the lock-in amplifier, the limited bandwidth applies.

In considering equivalent noise bandwidth, remember that the output meter is *an average responding meter calibrated to indicate the rms amplitude of a sinewave.* Gaussian noise, which has or has not been band-limited by filtering in the Signal and Output Channels, makes the meter indicate $\sqrt{\pi/2}$ times the actual rms value.

When the Q switch is set to 10% ENBW (bandpass mode), the Q of the tuned amplifier is not 15.7, as indicated above, but is instead 12.34. This lower Q is necessary to compensate the noise response of the ac voltmeter circuitry. If the voltmeter were a true rms responding voltmeter, then a Q of 15.7 would indeed be proper. With an average responding voltmeter circuit such as is employed in the Model 124A, exactly the same overall response is obtained with a Q of 12.34. In any case, 10% ENBW operation should prove suitable in most applications.

3.21 DYNAMIC RANGE

The TOTAL DYNAMIC RANGE of a lock-in amplifier is defined as the quotient of the maximum input that can be applied to the input without overload divided by the minimum discernible signal (MDS). Total Dynamic Range in turn is divided into two parts, each referenced to the input signal required to give full-scale output. The quotient of the amount of signal required for full-scale output divided by the minimum discernible signal is called the OUTPUT DYNAMIC RANGE. The quotient of the maximum input without overload (OVL) divided by the amount of signal required for full-scale output is called the DYNAMIC RESERVE of the lock-in amplifier. Thus, TOTAL DYNAMIC RANGE is simply the sum (logarithmic) of the OUTPUT DYNAMIC RANGE and the DYNAMIC RESERVE. All three are important in specifying the dynamic range characteristics of a lock-in amplifier, because, depending on where the division is made, the suitability of the lock-in amplifier to making a particular type of measurement can vary greatly. A lock-in amplifier with a Total Dynamic Range of 10^4 could have that Total Dynamic Range divided in several different ways. For example, it could have a Dynamic Reserve of 10^3 and an Output Dynamic Range of 10^1, in which case it would be well suited to processing very noisy signals but ill suited to processing small-amplitude noise-free signals. It could have a Dynamic Reserve of 10^2 and an Output Dynamic Range of 10^2, in which case it could still process moderately noisy signals and also be suitable for processing reasonably small noise-free signals. Finally, it could have a Dynamic Reserve of 10^1 and an Output Dynamic Range of 10^3, in which case its capability of processing a noisy signal would be severely restricted while its ability to process a small noise-free signal would be very good. Thus, the manner in which Dynamic Reserve is traded for Output Dynamic Range in a lock-in amplifier is one of the major factors involved in determining the suitability of the instrument for making a given measurement. In the case of the Model 124A, the operator has control of the dynamic tradeoff so that the dynamic range characteristics of the instrument can be optimized for the type of measurement at hand. The front-panel Function switch gives the operator the choice of LOW DRIFT, NORMAL, and HIGH DYNAMIC RANGE (Reserve) operation. Each position of the switch corresponds to a different division of the Total Dynamic Range into its Dynamic Reserve and Output Dynamic Range components.

The Dynamic Range characteristics of the Model 124A are illustrated in Figure III-18. Referring to the figure, note that two different OVL levels are indicated. The first of these, the PSD OVL level, defines where the Phase Sensitive Detector overloads relative to a full-scale input signal. It is this overload level that determines the maximum tolerable input signal when the instrument is operated in the FLAT mode. When the instrument is not being operated in the Flat mode, then the maximum tolerable input is extended to the PRE-PSD OVL for signals not in the passband of the characteristic selected (BANDPASS, NOTCH, HI PASS, LO PASS). Note that the PSD DYNAMIC RANGE of the instrument is 10^6 regardless of the selected dynamic tradeoff. Depending on the tradeoff selected, the PSD DYNAMIC RESERVE varies from 10^1 to 10^3 and the OUTPUT DYNAMIC RANGE varies from 10^3 to 10^5. The TOTAL DYNAMIC RANGE is 10^9 for LO DRIFT and NORMAL operation and 10^8 for HIGH DYNAMIC

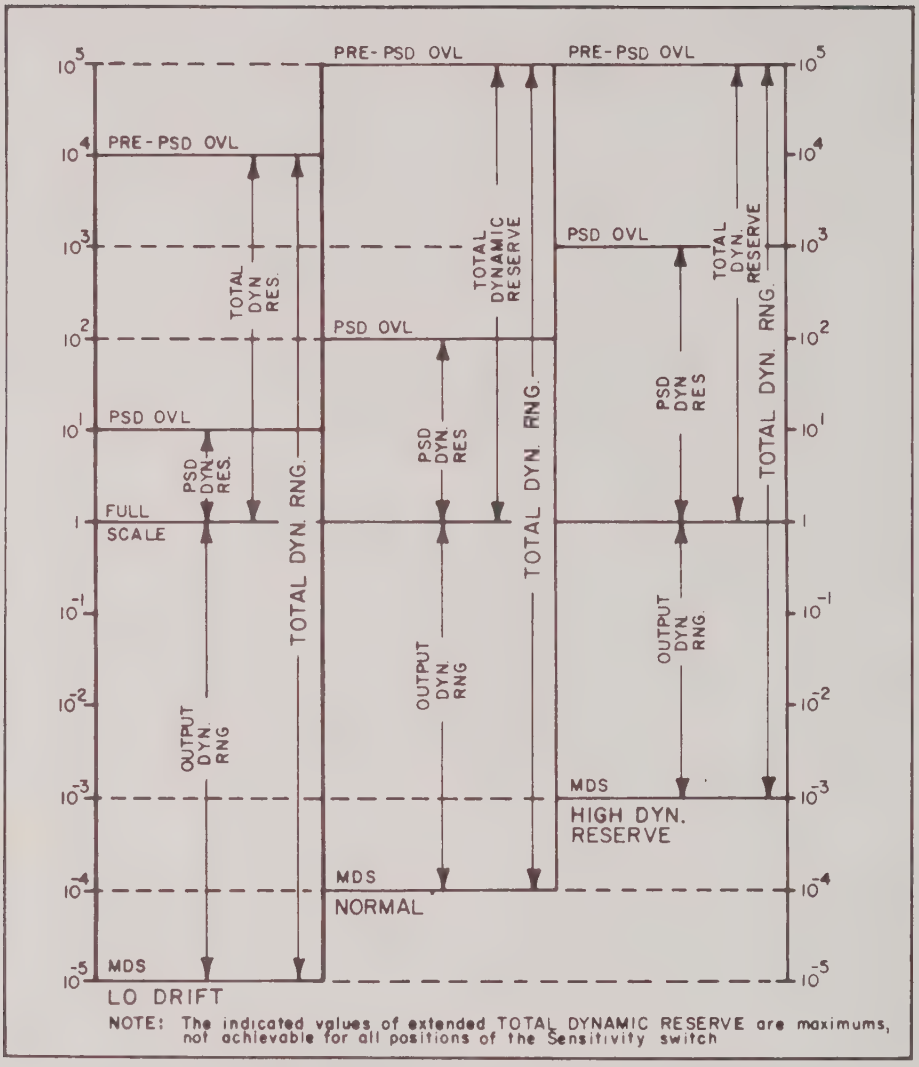


Figure III-18. DYNAMIC RANGE CHARACTERISTICS OF THE MODEL 124A

RANGE (Reserve) operation. In applications where the PSD OVL level is exceeded, the appropriate pre-PSD passband limiting is used to reduce the noise below the PSD OVL level so that the measurement can be made. Use of the Bandpass, Notch, Lo-Pass and Hi-Pass characteristics in this manner does not improve the overall achievable improvement in signal-to-noise ratio, but it does achieve a real improvement in Dynamic Reserve, and hence in Total Dynamic Range. In FLAT mode operation, the PSD DYNAMIC RANGE and the TOTAL DYNAMIC RANGE are the same.

It should be noted that the Dynamic Range characteristics shown in Figure III-18 are maximums, not applicable for all positions of the Sensitivity switch. In LO DRIFT operation, maximum Total Dynamic Range is attained with the Sensitivity switch set to 1 μV. Below 1 μV the Model 124A automatically transfers to NORMAL operation as described on page III-15. In NORMAL and HIGH DYNAMIC RANGE (Reserve) operation, maximum Total Dynamic Range is attained with the Sensitivity switch set to 100 nV. For all three Dynamic Tradeoff possibilities, as the Sensitivity switch is set to successively lower sensitivities, the Total Dynamic Range available is proportionally reduced. Situations could arise where the signal to be measured was of relatively high amplitude, and accompanied by enough noise and interference to require that the lock-in amplifier have a very wide Total Dynamic Range. Where this is the case, attenuators can be used ahead of the lock-in amplifier to reduce the signal plus interference sufficiently to take advantage of the inherently wide Total Dynamic Range of the Model 124A. It is almost inconceivable that a real measurement problem could exist that would require *all* of the Total Dynamic Range available with this instrument.

For a more detailed discussion of Dynamic Range, the reader is referred to the Appendix at the rear of the manual.

3.2J DYNAMIC OVER-RIDE

Although the front-panel FUNCTION switch allows the operator to select LO DRIFT, NORMAL, or HI DYNAMIC RANGE (Reserve) operation, it is important to understand that the instrument does not necessarily operate with the dynamic tradeoff selected. For certain positions of the Sensitivity switch, there is an "over-ride" action that determines the dynamic tradeoff *independent* of the setting of the Function switch, and it is the true or operating dynamic tradeoff that determines the overload and stability characteristics of the instrument. Thus, in using the tables that define the Overload and Output Stability characteristics of the Model 124A, the operator must always take care to read the data from the column corresponding to the true dynamic tradeoff, which may differ from that selected with the Function switch. The dynamic tradeoff obtained as a function of the setting of the Function and Sensitivity switches is as follows.

If the FUNCTION switch is set to: Then:

LO DRIFT AND ACVM: The unit operates in LO DRIFT except when the Sensitivity switch is set to 100 nV, 200 nV, or 500 nV. For those positions, the unit operates in the NORMAL mode and the NORMAL mode drift and noise tolerance specifications apply.

NORMAL: The unit operates in NORMAL with Sensitivity switch settings of 100 nV through 50 mV. With settings of 100 mV through 500 mV it transfers to LO DRIFT.

HI: The unit operates in HI with Sensitivity switch settings of 100 nV through 5 mV. With settings of 10 mV through 50 mV, it transfers to NORMAL, and with settings of 100 mV through 500 mV, it transfers to LO DRIFT.

Generally speaking, one should operate with the switch set to LO DRIFT to take advantage of the excellent output stability obtained with this setting (see Table III-1). However, if the noise level of the signal is high enough to overload the demodulator as explained in the following paragraphs, one can operate in either NORMAL or HI, in which the noise tolerance is increased by a factor of 10 or 100 respectively at the expense of degraded output stability.

DYNAMIC TRADEOFF	OUTPUT STABILITY	RMS OUTPUT NOISE VOLTAGE
LOW DRIFT	15 ppm/°C	10 ppm/$Hz^{1/2}$
NORMAL	100 ppm/°C	100 ppm/$Hz^{1/2}$
HI	1000 ppm/°C	1000 ppm/$Hz^{1/2}$

Table III-1. STABILITY AND OUTPUT NOISE AS A FUNCTION OF OPERATING DYNAMIC TRADEOFF

3.2K OVERLOAD

Depending on the nature of the input signal plus noise, and on the control settings, overload can occur at several different points in the instrument. All of the critical points are monitored so that an overload anywhere in the instrument will operate the OVERLOAD light. From an operator's point of view, the problem with regard to overload is not one of determining whether overload is taking place (the Overload indicator light performs this function automatically), but rather of determining the proper action to take to eliminate the overload. The appropriate remedial action in turn depends on where the instrument is overloading. Each overload "type" is considered in the following paragraphs.

(1) DC Amplifier Overload

Overload of the dc amplifiers is the easiest type of overload to identify and detect. Such overload is usually the result of the signal amplitude (independent of noise) being too large for the selected sensitivity. When this is the case, the panel meter indication exceeds full scale and the Overload light turns on. The solution is simply to select a sensitivity setting which yields an on-scale indication.

DC Amplifier overload can also be produced by a high-amplitude quadrature signal component, or by high amplitude spikes and other noise which may be reaching the Output Amplifier (the latter is particularly true when operating in the Flat mode). The test and solution is simply to increase the Time Constant setting. Except when operating with extremely noisy signals, a time constant of one second should suffice to eliminate quadrature and transient overload of the dc amplifiers.

(2) Demodulator Overload

Overload at the demodulator is also easily detected; one has only to monitor the signal at the Signal Monitor connector with an oscilloscope. If the signal exceeds 1.5 V pk, the demodulator is overloading. If the signal is less than 1.5 V pk, it is not.

There are several possible courses of action when faced with Mixer (demodulator) overload. First, one always has the option of operating the instrument with less sensitivity. Second, operating with a narrower bandwidth ahead of the Mixer may prove helpful. If the instrument is being operated in the Flat mode, and operation in the LOW-PASS, HIGH-PASS, or, better yet, the BANDPASS mode is possible, a considerable reduction in noise at the input to the Mixer may be achieved by making the transfer. If the instrument is operating in bandpass to begin with, increasing the Q will further decrease the bandwidth. If the interference is at a single frequency removed from the signal frequency, operating in the Notch mode with the signal channel tuned to the interference frequency may prove to be the best solution.

This third choice is to make a different tradeoff of dynamic reserve for output dynamic range. If the instrument is already operating with the Function switch set to HI, no improvement by this means is possible. If, however, it is operating with the switch set to NORMAL, a factor of ten reduction in the Mixer input signal amplitude can be obtained simply by setting the Function switch to HI. If the Function switch is set to LO DRIFT, a factor of 100 reduction in Mixer input signal amplitude is possible, the first factor of ten by setting the switch to NORMAL, the second by setting it to HI. In other words, in choosing the position for this switch, be sure to take the Mixer overload considerations into account as well as output drift requirements. Also, one should be aware of the dynamic over-ride transfers that take place for certain positions of the Sensitivity switch as explained earlier.

Note that the Function switch gives a simple "test" for Mixer overload as well, which, though not as definitive as monitoring the signal at the Signal Monitor jack, may still prove useful. If the Function switch is set to LO DRIFT, and the Overload light is on, and setting the Function switch to NORMAL or HI causes the Overload light to go out, the problem is obviously one of Mixer overload. Similarly, if the switch is set to NORMAL (Overload Indicator on), and setting it to HI causes it to go out, the same is true. It does not necessarily follow that failure of the light to extinguish when making this test means the Mixer is not overloading.

It may prove useful to know the maximum rms input sinewave signal that can be applied to the input of the instrument without overloading the Mixer as a function of the Sensitivity and Operating Dynamic Tradeoff. The input overload limits for both the mixer and the tuned amplifier are shown in Table III-2. By referring to this table, the operator can quickly determine the best way to process a given signal with the Model 124A. For example, suppose one had a signal of nominally 1 μV amplitude. Consider the possibilities for measuring this signal given different noise levels. From Table III-2, the mixer limit with a Sensitivity setting of one microvolt is 1.10 mV, 110 μV, or 11 μV, according to the dynamic tradeoff. What this means practically is that for noise levels below 11 μV, the signal could be measured in the FLAT mode with the Function switch set to LO DRIFT (Normal and Hi operation is, of course, also possible). If the noise level is greater than 11 μV, but less than 110 μV, LO DRIFT operation in the FLAT mode is not possible and the operator will have to either operate in NORMAL (or Hi) or narrow the noise bandwidth ahead of the Mixer by means of the tuned amplifier. This is usually best accomplished by operating in BANDPASS. By sufficient narrowing of the pre-demodulator bandwidth, LO DRIFT operation

MAXIMUM RMS SIGNAL INPUT

	MIXER LIMITS			TUNED AMP. LIMITS		
SENS.	H.D.R.	N	L.D.	H.D.R.	N	L.D.
100 nV	110 μV	11 μV	←	19.6 mV	19.6	←
200 nV	254 μV	25 μV	←	"	"	←
500 nV	635 μV	63 μV	←	"	"	←
1 μV	1.10 mV	110 μV	11 μV	"	"	19.6 mV
2 μV	2.54 mV	254 μV	25 μV	"	"	"
5 μV	6.35 mV	635 μV	63 μV	"	"	"
10 μV	11.0 mV	1.10 mV	110 μV	"	"	"
20 μV	19.3 mV	2.54 mV	254 μV	"	"	"
50 μV	"	6.35 mV	635 μV	"	"	"
100 μV	110 mV	11.0 mV	1.10 mV	196 mV	"	"
200 μV	193 mV	19.3 mV	2.54 mV	"	"	"
500 μV	"	"	6.35 mV	"	"	"
1 mV	720 mV	110 mV	11.0 mV	720 mV	196 mV	"
2 mV	"	193 mV	19.3 mV	"	"	"
5 mV	"	"	"	"	"	"
10 mV	→	720 mV	110 mV	→	720 mV	196 mV
20 mV	→	"	193 mV	→	"	196 mV
50 mV	→	"	"	→	"	196 mV
100 mV	→	→	720 mV	→	→	720 mV
200 mV	→	→	"	→	→	720 mV
500 mV	→	→	"	→	→	720 mV

Table III-2. MAXIMUM RMS INPUT LEVELS FOR MIXER AND TUNED AMPLIFIER OVERLOAD AS A FUNCTION OF SENSITIVITY AND OPERATING DYNAMIC RANGE

may still be possible even with the higher input noise level. If the noise exceeds 110 μV, but is less than 1.10 mV, FLAT mode operation is only possible with the Function switch set to HI, although NORMAL, or, conceivably, even LO DRIFT operation may still be possible by sufficiently narrowing the bandwidth ahead of the Mixer. If the Noise level exceeds 1.10 mV, then FLAT mode operation becomes impossible and the operator must narrow the bandwidth ahead of the Mixer. Finally, if the input noise exceeds 19.6 mV rms, the Selective Amplifier overloads and the signal cannot be measured with the Model 124A. The only possibilities in that case are to narrow the bandwidth ahead of the lock-in amplifier, or to attenuate ahead of the lock-in amplifier, whichever technique most conveniently reduces the input noise below the 19.6 mV limit.

As a general rule, it is most desirable to operate in LO DRIFT and with the bandwidth ahead of the Mixer at maximum (FLAT). If the noise levels are such as to force tradeoffs, the operator must decide which is better to give up first, output drift or a flat pre-mixer response, according to his individual requirements. Where the noise level is extremely high, there is no choice but to give up both. One way to approach the problem is to:

(a) Set the Function switch to ACVM to get an idea of the signal plus noise rms amplitude.

(b) Based on an estimate of the signal amplitude, determine the best combination of operating dynamic range and pre-detector bandwidth control from Table III-2. Be sure the Sensitivity setting considered is that appropriate to the expected signal level. For a noisy signal, the Sensitivity setting will be very much different from that used in the preceding step, where one was measuring the signal plus noise.

(c) Set the controls as determined in "b" and attempt the measurement. Some experimentation may be required to achieve the optimum control settings.

(3) Pre-Mixer Overload

Circuits ahead of the Mixer only overload when the input level exceeds the Tuned Amplifier Limits indicated in Table III-2. In any case, there is relatively little one can do at the Model 124A in the case of pre-mixer overload. The only action that might help is to reduce the sensitivity. For example, suppose one wished to measure a 500 μV signal accompanied by 400 mV of noise. From Table III-2 it is clear that the Tuned Amplifier would overload if the measurement were attempted on the 500 μV sensitivity range. However, by setting the Sensitivity switch to 1 mV, the overload tolerance is increased to 720 mV, and the measurement can be made. This technique is most useful when one is near an overload tolerance crossover point, as in the example just given. Should one be far from such a point, the likelihood of improving the situation in this matter becomes remote. For example suppose one had a 1 μV signal accompanied by 40 mV of noise. The nearest sensitivity position that could be used without overloading the Tuned Amplifier is 100 μV, where the signal would be only 1% of full scale. Given the high noise level, it would be very difficult to detect a 1% of full-scale signal, even if a very long time constant were used.

At no time can the input exceed 720 mV rms without overloading the instrument. With large input signals, this limit prevents one from processing very noisy signals. However, by inserting an attenuator ahead of the lock-in amplifier, large amplitude signals having poor signal-to-noise ratios can be measured.

3.2L OFFSET DUE TO NOISE

Because of imperfections in the Phase Sensitive Detector, large input noise levels can cause offsets to appear at the output. As shown in Figure III-19, these offsets are generally so small as to be negligible. Even with the extremely high 1000 times full scale noise levels which can be processed in high dynamic reserve operation, the offset is typically only 1%.

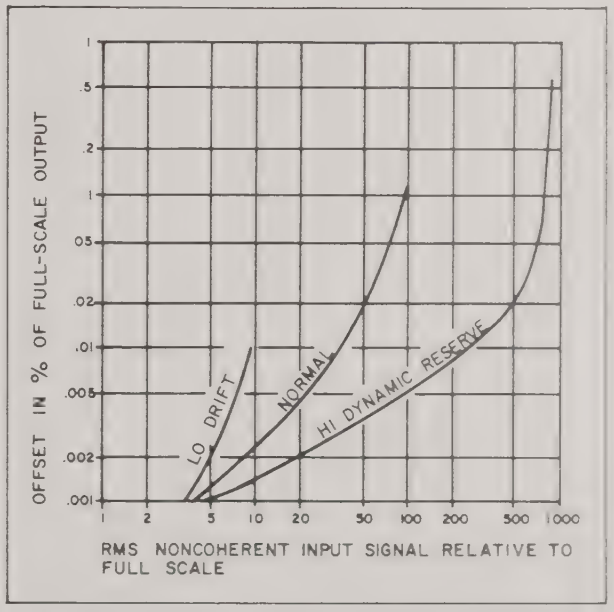


Figure III-19. TYPICAL OUTPUT OFFSET AS A FUNCTION OF INPUT NOISE

3.2M OVERLOAD RECOVERY

The frequency range of the Signal Channel, excluding the Preamplifier range, is 2 Hz to 210 kHz on the standard models, or 200 mHz to 210 kHz if requested upon ordering. Unless very low frequency response is really needed, we advise that units not be ordered with 200 mHz response because larger coupling capacitors make recovery time from overload much longer than for units with 2 Hz response. Typical maximum overload recovery time for units having 2 Hz LF response is 30 seconds, for 200 mHz units, 80 seconds.

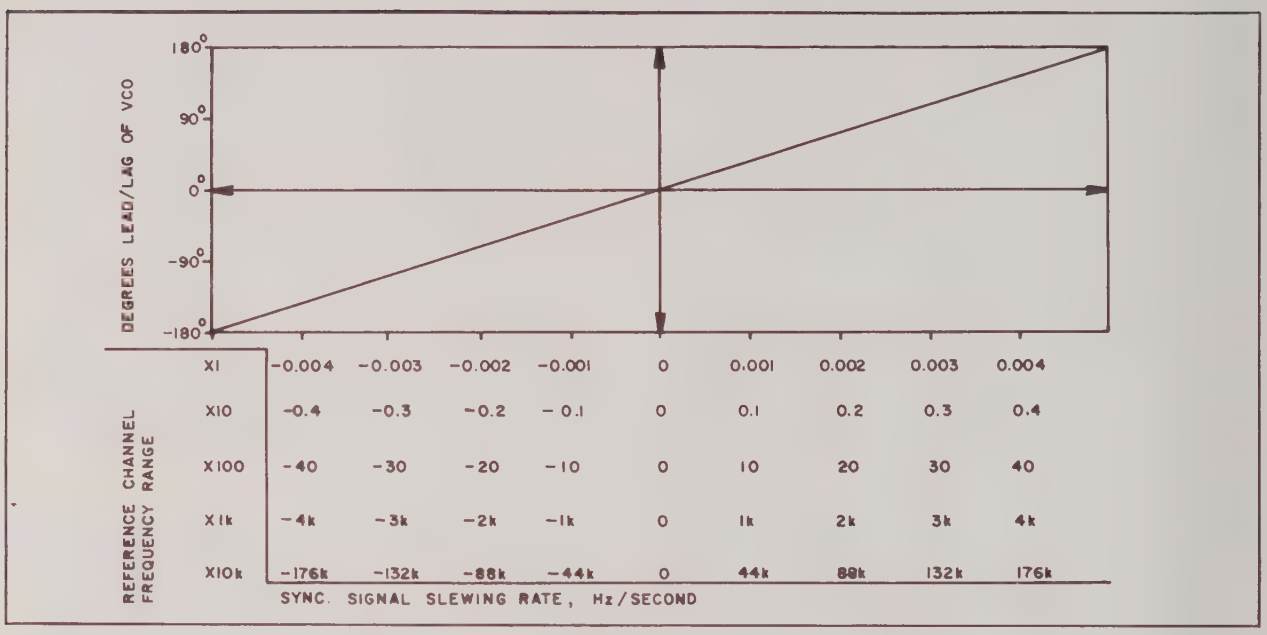


Figure III-20. TYPICAL REFERENCE OSCILLATOR SLEWING RATE

3.2N SIGNAL MONITOR

The output of the Signal Channel goes to the Synchronous Detector and to the Signal Monitor jack. The Signal Monitor output is useful for monitoring the signal after Signal Channel filtering; the operator can thereby improve his idea of how much noise is present ahead of the Mixer. In addition, this output makes the Model 124A usable as a straightforward Low Noise Tuned Amplifier, which can find many applications.

Notice from the Functional Block Diagram (Figure III-1) that the signal from the Signal Channel is attenuated before being applied to the Signal Monitor jack. This is because the amplification in the Signal Channel is such as to make the average-responding meter at the Detector output indicate the rms value of the detected signal. The attenuator at the Signal Monitor jack sets the signal to a more convenient level.

The Signal Monitor output impedance is 600 ohms. The output signal amplitude corresponding to a full-scale input depends on the dynamic range. In LO DRIFT, a full-scale input yields 100 mV rms out (sinewave in; sinewave out). In NORMAL, a full-scale input yields 10 mV out, and in HI DYNAMIC RANGE a full-scale input yields 1 mV rms out. These figures depend on the "true" operating dynamic range, which, as explained earlier, according to the selected sensitivity, can differ from the dynamic range selected with the Function switch.

3.3 REFERENCE CHANNEL OPERATION

3.3A SYNC INPUT/OUTPUT

In the External Sync mode, the Reference-Channel VCO automatically phase-locks to any kind of reference waveform at a frequency within the two decades listed in Table III-3, the only requirements being that the waveform cross its mean exactly twice each cycle, that it have a peak-to-peak amplitude of at least 100 mV, and that it be synchronized with the signal of interest. The positive-going zero crossing of the zero reference phase of the VCO sinewave is coincident with the positive-going zero crossing of the sync input waveform. The VCO can be made to lock to the second harmonic of the reference signal by placing the Mode switch in the Ext. f/2 position. Maximum sync input frequency in f/2 is 105 kHz. Sync input R = 1 megohm.

A Reference Unlock panel indicator lights when the VCO is out of sync with the reference signal.

Often, the frequency of the signal being detected is changing. As long as the reference maintains a fixed frequency and phase relationship to the signal, detection is no different than for fixed-frequency signals. However, it is necessary that the frequency does not change so fast that the oscillator cannot stay locked in. Figure III-20 provides slewing rate information for the oscillator.

Because of the rapidity of the External Reference circuitry response time, some care must be taken to assure that there are no "extra" mean crossings in the applied reference signal. If extra crossings should occur, the Model 124A will see them as bursts of some higher reference frequency, which the Reference Channel will attempt to follow, causing phase-lock loss and improper drive to the demodulator. There are three common ways in which problems of this nature occur. First, any noise accompanying the signal can cause multiple mean-crossings to occur in the region of the rise and fall of the reference signal. Thus the reference signal applied must be relatively clean. Frequently, moderately noisy signals can be cleaned up sufficiently for satisfactory operation by using a simple single-section

low-pass filter. Second, if the applied reference signal has very fast rise and fall times, it will be necessary to terminate the reference signal cable in its characteristic impedance to prevent ringing. Ringing of sufficient amplitude will upset the reference circuitry in the same manner as noise. The third way of introducing additional crossings is to allow interference from motors, light switches, or any spark generating equipment to contaminate the external reference signal. Use of a shielded cable to carry the reference signal will generally prove helpful. Also, whenever possible, the amplitude of the reference signal should be large relative to the amplitude of any possible interference. External reference signal levels on the order of several volts are advisable where interference could be a problem.

In either of the external modes, offset at the output of the Model 124A is less than 1 ppm for a shift in symmetry of the sync input signal from 0.1% to 99.9%. The response to even harmonics (Tables III-4 and III-5) is unchanged for this range of dissymetry.

In the Ext. and Ext. f/2 modes, one can greatly reduce the time that it takes the Oscillator to lock to the reference waveform if one knows in advance the frequency of the externally applied reference. To achieve phase lock in the minimum possible time, proceed as follows:

(1) Set the frequency controls to the desired frequency (for example, if the externally applied reference is 400 Hz and the desired mode is Ext., dial in 400 Hz; if the desired mode is Ext. f/2, dial in 800 Hz).

BAND	(FREQUENCY RANGE)	MAXIMUM TIME
X1	(200 mHz to 21 Hz)	15 min.
X10	(2 Hz to 210 Hz)	2 min.
X100	(20 Hz to 2.1 kHz)	10 sec.
X1K	(200 Hz to 21 kHz)	2 sec.
X10K	(2.1 kHz to 210 kHz)	2 sec.

Table III-3. MAXIMUM FREQUENCY ACQUISITION TIMES OF OSCILLATOR

(2) Set the Mode switch to INT. for a few (1 or 2) seconds. This presets the reference frequency so that the Oscillator does not have to search so far. Then switch back to Ext. or Ext. f/2, whichever is to be used.

In the Internal mode, a sinewave output from the Oscillator is provided for synchronizing the experiment. The Oscillator can free run at any desired frequency from 200 mHz to 210 kHz, and the frequency can be accurately set with the digital controls. It is possible to move the frequency above or below the dialed frequency by applying voltage to the rear-panel VCO Input connector. By setting the dials to zero, the frequency is determined entirely by the VCO input voltage, with 0 to 10 volts corresponding to the full range of the set band. The set band can be swept completely in 1 second.

The sync sinewave output level can be adjusted to any value from 0 to 10 V rms, with 8 intermediate calibrated levels. Output R = 600 ohms. This sinewave output is available even when the Oscillator is controlled by an external sync signal.

3.3B PHASE CONTROLS

The Phase controls set the phase of the synchronous detection process with respect to the phase of the sync signal. The operator adjusts the Phase Controls for maximum meter indication. At maximum, the meter reading represents the detected signal amplitude, and the phase of the signal can be read from the phase dials.

The ease with which the fine phase control can be adjusted depends on how steadily the meter holds, and this depends in turn on the steadiness of the signal amplitude and the final noise level. The final noise level depends on the original noise level and how well the Signal Channel and output filters are set.

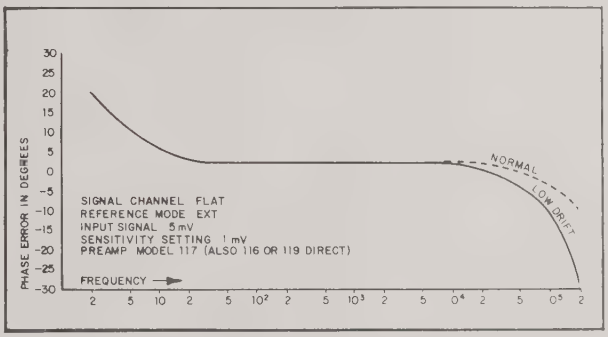
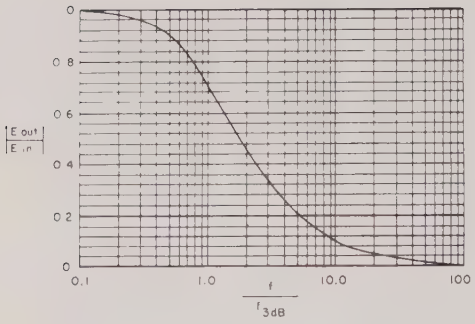


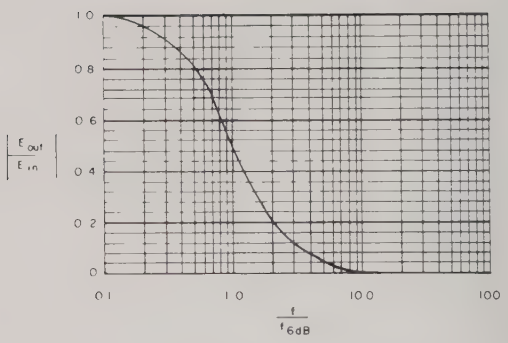
Figure III-21. NET PHASE DIFFERENCE BETWEEN SIGNAL AND REFERENCE CHANNELS AS A FUNCTION OF FREQUENCY

When adjusting the phase controls for maximum, the amplitude peak varies as the sine of the phase adjustment; i.e., the fine peak adjustment is relatively flat, and a small amount of residual noise could make the peak hard to find. One way to get around this is to adjust the controls for as close to a peak as possible, then change the Phase Quadrature control setting by 90° and adjust for zero. This quadrature zero varies as the cosine of the phase adjustment, and is, therefore, more sensitive than the peak adjustment. The Phase Quadrature control can then be switched back to the in-phase quadrant and the reading taken. If the output signal is very clean, adjusting for the peak, rather than for quadrature null, may be more accurate because the effects of phase control error are avoided.

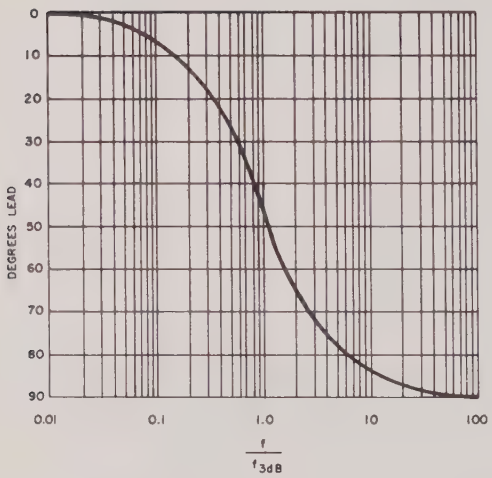
The ten-turn potentiometer, which spans 100°, together with the quadrant switch, allows the entire 360° to be covered with 0.1° resolution. The linearity is within ±2 from 2 Hz to 21 kHz and ±5° above 21 kHz. Remember that the actual difference of phase between the signal and reference channels can be in excess of the phase-shifter accuracy, depending on the filter settings and also on the frequency. Special procedures for accurately measuring the phase are given in Subsection 3.9.



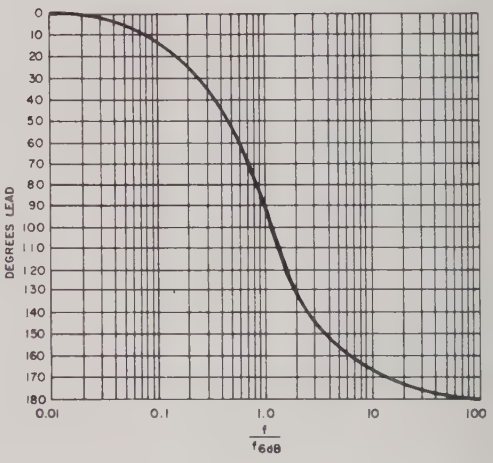
6 dB/Oct. Amplitude Transfer



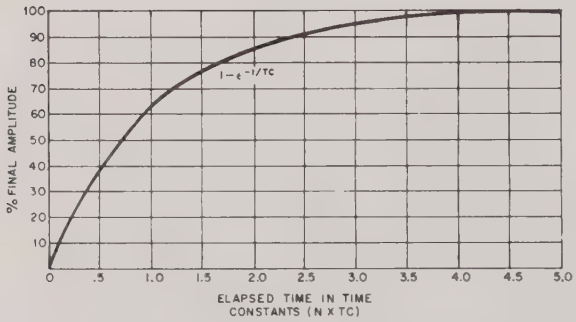
12 dB/Oct. Amplitude Transfer



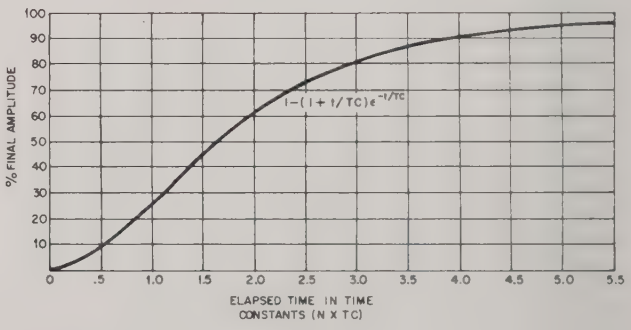
6 dB/Oct. Phase Transfer



12 dB/Oct. Phase Transfer



6 dB/Oct. Step-Function Response



12 dB/Oct. Step-Function Response

Figure III-22. OUTPUT FILTER TRANSFER FUNCTIONS

NOTE: When operating the Signal Channel in the Bandpass mode, particularly with high Q, the frequency dials must also be fine-adjusted for peak meter reading. Otherwise, the filter rolloff will attenuate the signal and also cause a phase error. In all circumstances, attenuation and phase shift in the Signal Channel must be minimized or accounted for.

3.4 OUTPUT CHANNEL OPERATION

3.4A FILTER TIME CONSTANT

The amplified and filtered signal is synchronously detected with respect to the phase-shifted Reference Channel signal, and the detected signal is applied to the RC low-pass filter. Either a one section or a two section filter may be selected with the center knob of the Time Constant switch.

The one section filter has a 6 dB/octave rolloff characteristic and an equivalent noise bandwidth of 1/4TC. The 3 dB-down point on the frequency axis is $1/2\pi TC$. In the time domain, it has a step-function response of $1-e^{-t/TC}$. The rise time from 10% of full amplitude to 90% of full amplitude is 2.2TC seconds, and from 0 to 95% is 3TC seconds. With the TC switch set to 300 s, the equivalent noise bandwidth is $\cong 800\ \mu Hz$.

The two section filter has a 12 dB/octave rolloff characteristic and equivalent noise bandwidth of 1/8TC. The 6 dB-down point on the frequency axis is $1/2\pi TC$. In the time domain, it has a step-function response of

$$1 - (1 + t/TC)e^{-t/TC}$$

The rise time from 10% of full amplitude to 90% of full amplitude is 3.3TC seconds, and from 0 to 95% is 4.8TC seconds.

If the operator requires a time constant greater than 300 seconds, he can place the Time Constant switch in the Ext. position and connect a pair of capacitors of equal value between pins 8-9 and 10-11 of the rear-panel octal socket. To determine the time constant for this external mode, multiply the single capacitor value (in Farads) by 30 megohms. The external capacitors should be low-leakage film types (mylar, polycarbonate, polystyrene, teflon) rated at 25 V or higher.

The Time Constant should be set so that the output noise, either as read on the panel meter or on the external monitor, is reduced to an acceptable level. If the signal amplitude is steady (independent of noise), a fairly long time constant can be used because the lag time in setting the phase controls can usually be tolerated. However, if the signal varies over a period of time, and the operator wants to observe the variations, a shorter time constant must be used at the expense of greater noise. Sometimes, for the latter case, the Signal Channel filter can be readjusted for less noise after operating parameters are better established.

When operating the instrument as a wideband ac voltmeter (by placing the function switch in the ACVM position), the Time Constant switch should be set to dampen the meter.

3.4B OFFSET CONTROLS

The ten-turn dial and its associated polarity switch allow calibrated offsets of up to ten times full scale to be applied. Two applications for this feature are that it allows small amplitude variations in a signal to be expanded and examined in detail, and that it allows a signal amplitude to be read with greater resolution than is possible with the panel meter alone. For example, suppose one had a meter indication to the right. To read the amplitude with the greatest possible resolution, the polarity switch would be set to "+" and the dial adjusted for "null" at which time the signal amplitude could be read directly from the dial.

The following example illustrates use of the Zero Suppress feature to "expand" signal amplitude variations. Suppose one had a 70 μV signal. Assuming this signal were measured on the 100 μV sensitivity range, the resulting meter indication would be 70% of full scale. To examine small variations in this signal, one would first set the polarity switch to "+" (assume initial meter indication was to the right), followed by adjusting the dial for null. The dial setting required would be 0.70 and the meter sensitivity would be $\pm 100\ \mu V$ with respect to the 70 μV ambient level. A recorder connected to the output would allow the amplitude variations as a function of some experimental parameter to be recorded.

Because the range of the Offset dial extends to ten times full scale, the measurement can be modified slightly so that the amplitude variations are greatly expanded. In the example at hand, the Sensitivity switch could be set to 10 μV. Because the signal amplitude, 70 μV, is less than ten times the selected sensitivity ($10 \times 10\ \mu V = 100\ \mu V$), it is within range of the offset dial. If the dial were adjusted for null (setting of 7.00), the meter range would then be ± 10 μV full scale with respect to the 70 μV ambient signal level.

3.5 HARMONIC RESPONSE

The Synchronous Detector responds to signals which are harmonically related to, and synchronized with, the fundamental. The harmonic response is less than the fundamental response, but still may be large enough to cause significant errors in the fundamental measurement.

Remembering that synchronous detection has features similar to full-wave rectification, one would see immediately from symmetry considerations that the response to even harmonics is different from the response to odd harmonics. For odd harmonics, the detector response relative to the fundamental response is simply 1/n, where n is the number of the harmonic. For example, the third harmonic response is 1/3 the fundamental response, the fifth harmonic response is 1/5 the fundamental response, etc.

Theoretically, the Synchronous Detector should have no response at all to even harmonics. However, the reference waveform is not perfectly symmetrical, causing a small

response. The even harmonic response due to dissymmetry can be expressed as:

$$\text{Response Relative to Fundamental} = \frac{\sin n\pi(1 + e)/2}{n}$$

where n is the number of the fundamental and e is the fractional departure of the half period from the actual half period of the reference waveform. If n is even and ne is very much smaller than unity, the expression simplifies to, approximately, $\pi e/2$. In other words, the response to all even harmonics is about the same and is determined by the symmetry of the reference signal at the Phase Sensitive Detector.

In general, a given half-cycle of the reference signal will be within 0.3% of half the period, giving 0.003 as the value of e. Inserting this into the formula, one obtains 0.5%, a good working value for the response to even harmonics. Actual symmetry error varies with the phase setting, so that the response to even harmonics can be less than 0.5%.

The harmonic sensitivity of the Synchronous Detector is one reason for minimizing the passband of the Signal Channel. The portion of overall output due to harmonics will be reduced by the attenuation factor the Signal Channel provides at those frequencies. Transfer curves in Figures III-14 through III-17 should be referred to. Table III-4 lists typical measured synchronous detector responses to harmonics. Table III-5 lists overall responses to these same harmonics if operating in the Bandpass mode with a Q of 10.

Harmonic	0°	90°	180°	270°
2nd	0.15%	0.5%	0.2%	0.5%
3rd	35%	35%	35%	35%
4th	0.13%	0.55%	0.25%	0.7%
5th	15%	15%	15%	15%

Table III-4. TYPICAL HARMONIC RESPONSE OPERATING IN THE FLAT MODE

Harmonic	0°	90°	180°	270°
2nd	0.017%	0.017%	0.008%	0.04%
3rd	1.2%	1.2%	1.2%	1.2%
4th	0.0035%	0.014%	0.006%	0.017%
5th	0.35%	0.35%	0.35%	0.35%

Table III-5. TYPICAL HARMONIC RESPONSE OPERATING IN BANDPASS MODE WITH Q = 10

These considerations and examples assume the worst possible phase relationship between the harmonic and the reference signal. Practically encountered phase relationships are not usually the worst case, and the true error will be smaller than the computed error.

A related problem is that of errors resulting from sub-harmonic components of the input signal. Sub-harmonic signals do not *directly* contribute to the output indication; that is, the Detector does not respond to them. However, if a sub-harmonic is distorted in the Signal Channel, output errors can be introduced. This is because distortion results in higher harmonics being generated, to which the Detector is sensitive, especially the harmonic at the primary frequency of detection.

Sub-harmonics are not generally a significant component of the input signal, except, perhaps, when the Reference Channel is operated in the Ext. f/2 mode. In this situation, and ones similar to it, the second harmonic is regarded as the fundamental for the Signal Channel, and the original sync input fundamental is regarded as a large-amplitude sub-harmonic. Care must be exercised to not distort the fundamental, although it can be attenuated by filtering in the Signal Channel. Distortion in the Signal Channel, other than due to OVLD clipping, occurs mostly in the Preamplifier.

3.6 SENSITIVITY AND NOTCH CALIBRATION

The accuracy of the sensitivity calibration in the Bandpass mode requires that the notch be properly adjusted. Therefore, each time the sensitivity is calibrated, it is wise to first make the fine notch adjustment. In HIGH PASS and LOW PASS, the notch adjustment has very little effect because a Q of one is normally used in these modes. In FLAT, the adjustment has no effect at all. The procedures given here are combined procedures, first for adjusting the notch and then for calibrating the sensitivity.

The fine Notch adjustment should be made for exactly zero center-frequency signal transfer through the Signal Channel when in the Notch mode. However, because the Calibrator output is a square wave, containing harmonics, it would be difficult to use this signal for making the fine notch adjustment in the Notch mode; the harmonics would disallow an output zero, and adjusting for a definite value would be difficult. The Reference Channel's sinewave output would be better, but it is not pure enough to allow for an accurate setting. One good way to adjust the notch in the Notch mode is by using a separate high-purity sinewave source.

An alternative way to make the fine notch adjustment is to adjust the Notch Adj. control for proper center-frequency signal transfer through the Signal Channel when in the Bandpass mode. If the adjustment is off, changing the Q will change the gain at the center frequency. Making the fine notch adjustment is easy if this last fact is made use of; i.e., the fine notch adjustment should be made such that a change of Q does not change the gain at the center frequency.

The latter method has the merit of allowing a less-pure waveform to be used because the harmonics are eliminated in the Bandpass mode. If Q's of 100 and 50 are used, the Calibrator waveform can be used for an adjustment accuracy within a few tenths of a percent. If the Reference

Channel's sinewave is used, an accuracy of better than a tenth of a percent can be achieved. The following procedure uses the Calibrator output because it is more convenient to continue on into the sensitivity adjustment. The operator ought to be able to adapt this procedure to using the reference sinewave if he requires more accuracy.

Make the following preliminary control settings.

(1) Reference Channel

The Frequency switches should be set to or near the frequency which is to be used or expected when operating with the experiment, and the mode switch should be set to Internal.

The Phase quadrant selector should be set to 270°, and the fine phase control set to 90°. Note that this adds up to 0°; setting this way allows the fine control's overlap to adjust the phase through 0°, when adjusting for maximum meter indication in the steps that follow.

(2) Preamplifier

Operate single-ended direct. Connect the Calibrator output to the Preamp input. (Use a short cable, RG-58/U or RG-59/U having BNC connectors at both ends.)

(3) Output Channel

The Time Constant switch should be set such that the signal driving the meter is well filtered. However, too long a time constant will make the adjustment time too long. Typically, if the operating frequency were 400 Hz a good time constant setting is 100 ms.

The Zero Offset toggle should be set to the neutral off position, and the Function switch set to the PSD position to be used in the experiment. (The gain error introduced when switching from one PSD position to another is very small, about 0.5% maximum.)

(4) Signal Channel

The Sensitivity switch should be set to the intended operating position. However, if one of the nV ranges is used, internal noise can cause the meter reading to waver too much for making an accurate adjustment. It is better, therefore, to use one of the μV or mV positions. The gain error when switching back down to the nV range for operation will be less than the adjustment error if adjusted in a nV range. The Calibrator switch should be set to the same level as the Sensitivity switch.

The Frequency switches should be set to exactly the same settings as the Reference Channel frequency switches.

The Mode switch should be set to the Bandpass position.

The Q switch should be set to 50.

Proceed to make the Notch adjustment as follows.

(1) Adjust the Signal Channel frequency control and fine Phase control for maximum meter indication.

(2) Change the Q switch setting to 100; if the meter indication changes, adjust the Notch Adj. screwdriver control to minimize the change. Continue switching the Q back and forth between 50 and 100 and adjusting the Notch Adj. potentiometer for no change in the meter reading.

Now that the Notch is adjusted, before adjusting the fine sensitivity control, make the following changes in the control settings:

The Calibrator output is a square wave having its *fundamental* rms component as indicated for each switch position. Therefore, when calibrating the sensitivity, the Mode switch should be set to the Bandpass position and the Q switch set at 10 or higher (if the Q to be used when operating is higher than 10, calibrate with that Q setting). All other controls should be left as set for the notch adjustment.

An attenuator, packaged in a small box having male and female BNC connectors, is provided with each Model 116 and 119. This attenuator attenuates 100:1 so that the Model 124A can be calibrated in the transformer mode. The attenuator output Z is 1 ohm. If the transformer mode is to be used, connect the male BNC connector of the attenuator directly to the input BNC, and connect the calibrator output to the attenuator's female jack with a short cable. It is important that, if another source is used for calibration, the source presents an impedance of exactly 50 ohms to the attenuator.

After these settings are made, the actual procedure is simple: Adjust the fine Phase control for maximum meter indication. Then adjust the Sensitivity screwdriver control for an accurate full-scale meter indication.

3.7 AC VOLTMETER OPERATION

With the Function switch in the ACVM position, the Signal Channel output is used to operate the Synchronous Detector. This makes the Model 124 operate as an ac rms voltmeter.

If the signal driving the synchronous detector is a clean sinewave, the rms voltage indication will be very accurate. The filter in the Signal Channel can be used to clean up a waveform if necessary. If the waveform is not clean, however, the meter reading will still be within 10 or 12 percent of the wideband rms amplitude of the input signal plus noise. This wideband capability is very useful for wideband noise and complex waveform measurements.

As in lock-in operation, the Time Constant is used to smooth the meter indication. However, bear in mind that

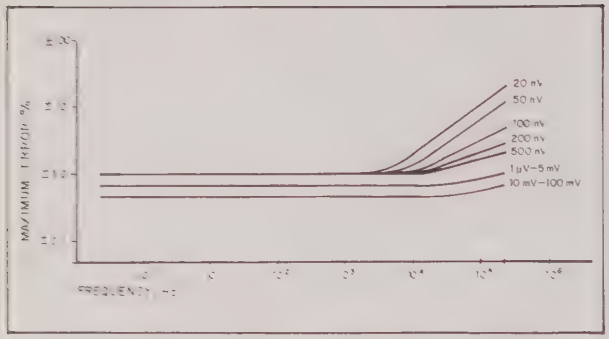


Figure III-23. TYPICAL CALIBRATION ACCURACY

both the signal and the noise contribute to the dc output of the detector in ACVM operation. The signal-to-noise ratio is not improved. Instead, the signal *plus* noise is measured, with the time constant serving solely to smooth the output indication.

It should be noted that, in ACVM operation, the Model 124A is effectively in the LO DRIFT mode, except for the 100 nV, 200 nV, or 500 nV sensitivity settings, where the unit operates in the NORMAL mode. Consequently, for all but the three "NORMAL mode" ranges, the drift and overload characteristics of the LO DRIFT mode apply.

3.8 DIGITAL PANEL METER MODIFICATION (1241/98)

If requested upon purchasing, a 3½ digit Nixie* display may be installed instead of the panel meter. This display provides direct numerical readout of the output, and the corresponding digital logic is available at a rear-panel connector. This logic is well suited for sending Lock-In output information to a computer via a Model 131 Instrument/Computer Interface System.

In reading the display, the numerals correspond directly to the signal voltage. However, some care must be exercised in interpreting the decimal indication. On any "1" range, 1 μV, 10 μV, 100 μV, 1 mV, 10 mV, 100 mV, etc., at full scale output the meter will read 1.000. Above full scale the meter will follow until overload occurs. On "2" ranges, however, at full scale the meter will "try" to read 2.000, but will instead read 1 BLANK, i.e., the decimal and three right-hand numerals will not illuminate at full scale. On "5" ranges, at full scale the meter will read .500, and above full scale the meter will follow until overload occurs. The voltage output at full scale is ±10 V on all ranges. No polarity symbol is displayed for positive readings. A "–" sign is displayed for a negative reading.

The information displayed on the digital meter is provided in binary coded decimal form at connector J7 at the rear. Table III-6 identifies the pins at which this information is provided and gives the output levels. Positive logic is employed; a "1" is +3.5 V ±1 V and a "0" is 0.2 V ±0.2 V. All digital output signals are capable of sinking 5 mA at the lower logic level and sourcing 0.1 mA at the upper logic level.

During normal operation, the digital meter is triggered internally at a rate of approximately 10 times per second. Other internal trigger rates can be obtained by changing the value of the resistor (nominally 300 kΩ) connected between pins 1 and 15 of DJ-2 (upper connector on Digital Panel meter). With this resistor removed, the rate is reduced to twice a second. There may be occasions where it is advantageous to trigger externally, such as to facilitate operation of the Model 124A in conjunction with other signal processing equipment which may be monitoring the digital output. Considerations that govern external triggering are that the internal trigger must be inhibited and that the proper external trigger must be applied. Internal triggering is inhibited by grounding pin 23 ($\overline{\text{BUSY}}$) of the digital output connector. The external trigger is applied to pin 20; it must be a logic one that goes to logic zero for at least one and a half microseconds (but for less than two milliseconds). This unit resets on the negative-going transition; conversion commences on the positive-going transition. The maximum allowable external trigger rate is 60 Hz.

When the Model 124A is being operated internally triggered (the usual case), it is important that signals be provided to indicate when a conversion is in progress (or not in progress as the case may be) if the Model 124A is to be successfully incorporated into a larger digital system. Three different signals are provided for this purpose. The first is the $\overline{\text{END OF CONVERSION}}$ level at pin 14 of the Digital Output connector. This output is up (nominally +3.5 V) while a conversion is in progress, and down for the full duration of the display plus reset time. The second signal, provided at pin 19 of the Digital Output connector, is a 75 μs logic one pulse, CONVERSION COMPLETE, generated at the end of the conversion period. The third signal, provided at pin 18, is the inverse, $\overline{\text{CONVERSION COMPLETE}}$, also 75 μs but at logic 0.

In the case of an instrument equipped with both the Remote Programming Option and Digital Panel Meter option, there are some special considerations that must be observed for proper operation. In units equipped with the Digital Panel Meter option alone, switching controlled by the front-panel Sensitivity switch sets the digital display sensitivity. In the case of units equipped with the Remote Programming option, the Sensitivity switch is rendered ineffective when the sensitivity is being controlled remotely, and control over the digital display sensitivity is lost.

To solve this problem, an additional toggle switch has been added to the rear panel of instruments equipped with both options. This switch has two positions, NORM, and D.P.M. 1.000. For operation with the remote-programming option *inactive* (this option is controlled by a rear-panel pushbutton), the switch should be set to NORM, in which position the display functions exactly as described in Subsection 3.8. In remote-programmed operation, the switch should be set to D.P.M. 1.000, in which position the digital panel meter indicates the input signal level as a fraction of *full scale*, independent of the selected sensitiv-

*Nixie is a trademark of Burroughs Corporation.

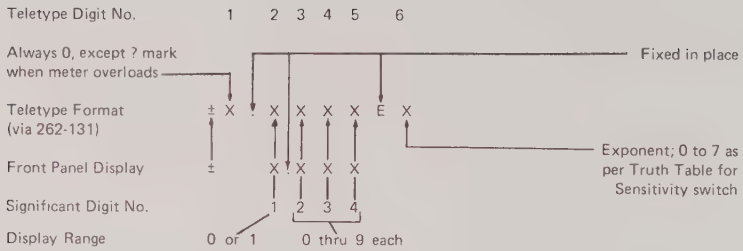
ity. A full-scale input gives a display indication of 1.000, independent of whether the programmed sensitivity is 1 mV, 2 mV, 5 mV, or some other value. Similarly, with a full-scale input applied the BCD output will be 1.000 and the recorder output will be 10 V. The display indication and output levels are proportionally less with less than full-scale inputs. For example, if the programmed sensitivity is 200 mV, and a 100 mV signal is applied, the digital display will indicate 0.500 (half scale), the BCD output will be 0.500, and the recorder output will be 5 V.

Model 124A J7 Pin #	Function	Function When Applied to Model 131 System Model 262
1	Polarity (Logic 1 = +)	1 = Polarity +
2	+4 volts dc	Not used
3	Overload Output (Logic 1 = Overload)	Digit 1, A
4	Overload Output (Logic 1 = Overload)	Digit 1, C
5	DVM Most Significant Digit, A	Digit 2, A
6	Digital Ground	Ground
7	DVM 2nd Most Significant Digit, A	Digit 3, A
8	DVM 2nd Most Significant Digit, C	Digit 3, C
9	DVM 3rd Most Significant Digit, A	Digit 4, A
10	DVM 3rd Most Significant Digit, C	Digit 4, C
11	DVM Least Significant Digit, A	Digit 5, A
12	DVM Least Significant Digit, C	Digit 5, C
13	Spare	
14	$\overline{\text{END OF CONVERSION}}$	
15	Spare	
16	Spare	
17	DVM Least Significant Digit, D	Digit 5, D
18	$\overline{\text{CONVERSION COMPLETE}}$	$\overline{\text{EXECUTE}}$
19	CONVERSION COMPLETE	Not used
20	Ext. Trigger Input	Not used
21	Spare	
22	Spare	
23	NOT BUSY input (Output data will remain fixed when this line is at logic 0. Must be 1 or open for conversions to continue)	$\overline{\text{BUSY}}$ (from Model 262)
24	Sensitivity Switch Position A (see Truth Table)	Digit 6, A
25	Sensitivity Switch Position C (see Truth Table)	Digit 6, C
26	Sensitivity Switch Position B (see Truth Table)	Digit 6, B
27	Digital Ground	Digit 6, D
28	DVM Overload Output (Logic 1 = Overload)	Digit 1, B
29	DVM Overload Output (Logic 1 = Overload)	Digit 1, D
30	Digital Ground	Digit 2, B
31	Digital Ground	Digit 2, D
32	DVM 2nd Most Significant Digit, B	Digit 3, B
33	DVM 2nd Most Significant Digit, D	Digit 3, D
34	DVM 3rd Most Significant Digit, B	Digit 4, B
35	DVM 3rd Most Significant Digit, D	Digit 4, D
36	DVM Lease Significant Digit, B	Digit 5, B

NOTE: Logic 1 = +3.5 V ±1 V, Logic 0 = +0.2 V ±0.2 V.

The DVM used in the Model 124A has a "3½" (4 digits) Nixie tube display. Each digit is represented at the rear panel connector in Binary Coded Decimal (BCD) format. The Most Significant Digit is the leftmost of the [illegible]. The Teletype digit #1 (? for overload) is 1 digit to the left of the Most Significant Digit. The notation A, B, C, & D [illegible] notation above refers to column headings of the truth table. The value of each of these outputs when at a logic 1 is 1, 2, 4, & 8, respectively. For each digit, the A, B, C, & D outputs, taken together, represent a number (0 to 9) in BCD format.

Figure III-6A. DIGITAL OUTPUT PIN ASSIGNMENTS


Teletype Digit No. 1 2 3 4 5 6
Always 0, except ? mark when meter overloads
Fixed in place
Teletype Format (via 262-131)
± X . X X X X E X
Front Panel Display
± X . X X X X
Exponent; 0 to 7 as per Truth Table for Sensitivity switch
Significant Digit No.
1 2 3 4
Display Range
0 or 1
0 thru 9 each

Printout is in microvolts

Digital Output for Each Display Figure:

Display Shows	Binary Coded Decimal Digital Output			
	8 (D)	4 (C)	2 (B)	1 (A)
0	0	0	0	0
1	0	0	0	1
2	0	0	1	0
3	0	0	1	1
4	0	1	0	0
5	0	1	0	1
6	0	1	1	0
7	0	1	1	1
8	1	0	0	0
9	1	0	0	1

Digital Output for Each Switch Setting:

Exp.	Switch Setting	BCD Digital Output		
		4 (C)	2 (B)	1 (A)
7	500 mV	1	1	1
6	200 mV	1	1	0
	100 mV	1	1	0
	50 mV	1	1	0
5	20 mV	1	0	1
	10 mV	1	0	1
	5 mV	1	0	1
4	2 mV	1	0	0
	1 mV	1	0	0
	500 μV	1	0	0
3	200 μV	0	1	1
	100 μV	0	1	1
	50 μV	0	1	1
2	20 μV	0	1	0
	10 μV	0	1	0
	5 μV	0	1	0
1	2 μV	0	0	1
	1 μV	0	0	1
	500 nV	0	0	1
0	200 nV	0	0	0
	100 nV	0	0	0

NOTE: Princeton Applied Research Corporation manufactures a cable suitable for interconnecting a Model 124A and a Model 262 Teleprinter/System Interface Module (part of the Model 131 Instrument/Computer Interface System). The part number of this cable is 6020-0023-06. A schematic drawing of this cable is included in Section VII on page VII-31.

Table III-6B. DIGITAL OUTPUT TRUTH TABLES

3.9 PHASE MEASUREMENTS

To measure phase with a Model 124A, the Phase controls are adjusted to obtain a *positive* peak indication, the same as for an amplitude measurement. When positive peak is achieved, the Phase controls will indicate the number of degrees by which the input signal *leads* the applied reference. If the angle is greater than 180°, it may be more convenient to subtract the indicated angle from 360° and state the difference as the angle by which the input signal lags the reference. **NOTE**: Any internal phase shifts, such as might be introduced by the signal amplifiers, must be taken into account for accurate measurements.

Sometimes, after the lock-in amplifier has been peaked up, a change in an experimental parameter will cause a phase shift and a resultant loss in peak indication. If this happens, it may be of interest to know whether the shift was a lead or a lag. To determine the direction of a phase shift, simply re-adjust the M124A Phase controls as required to restore the peak indication, and while so doing, note whether the new phase setting is higher or lower than the old one. If the new phase setting is higher, the phase shift of the signal relative to the reference was in the leading direction. If the new phase setting is lower, then the shift was in the lagging direction.

As mentioned previously, when the phase controls are adjusted for peak meter reading, they indicate the phase of the signal with respect to the reference signal. However, a more accurate determination can be made by taking advantage of the greater quadrature adjustment sensitivity. Also, the phase shift differences between the Reference Channel and the Signal Channel must be accounted for if an accurate phase determination is required. When adjusting for the peak, the meter reading varies around the peak as the sine of the phase angle for small errors. If the phase is adjusted for quadrature null instead, the meter reading varies around null as the cosine of the phase angle for small errors. Therefore, for any small number of degrees change of the Phase vernier while adjusting for a quadrature null, the meter reading changes much more than for the same vernier change while adjusting for an in-phase peak. Two high-sensitivity procedures follow.

Procedure #1

This procedure is relatively simple, and can be used with signals that vary in amplitude (independent of noise).

(1) Measure the amplitude of the signal in the normal manner, so that the controls are initially optimized for time constant, dynamic range, Signal Channel filter settings, etc. Use the Bandpass mode and high Q, if possible (frequency constant), to eliminate the effects of harmonics on making the null settings below. If the frequency is changing, it would be best to use a wide bandwidth and avoid using a transformer input, because internal phase variations as a function of frequency could not be accounted for.

(2) Disconnect the signal from the Preamplifier, and connect the sinewave from the Reference Channel Out jack to the Preamplifier's A input. If this output is already being used for synchronizing the experiment, use a "T" connector. The amplitude must be low enough so that the Signal Channel is not overloaded. The phase of this sinewave is going to be used for zero reference. If a clean sinewave having a phase more suitable for zero reference is available, use it.

If a transformer input is used, it is important that the Reference output impedance be made to look the same as that of the signal source. An appropriately designed attenuator can generally be used to achieve this goal. Otherwise, phase measurement errors will be introduced.

(3) Set the Preamplifier Mode switch to "A". (The Transformer/Direct switch should be set as appropriate for the intended input coupling.)

(4) Set the Zero Offset switch to OFF (center position).

(5) Adjust the Phase controls for an exact meter null. Increase the Signal Channel sensitivity as much as possible without overload while making this adjustment.

(6) Record the Phase dial setting (phase zero).

(7) Disconnect the Reference Channel sinewave from the Preamplifier, and reconnect the Reference Channel sinewave from the Preamplifier, and reconnect the signal to be measured.

(8) Adjust the Phase controls for exact meter null. Increase the Signal Channel sensitivity as much as possible without overload while making this adjustment.

(9) The difference between the zero reference phase recorded in step (6) and the phase set in step (8) is the accurate phase of the signal with respect to the reference signal.

Procedure #2

This procedure is more complicated, but it has the advantage of providing a voltage output (& digital output if the optional digital meter is installed) proportional to the cosine of the phase angle, which can be sent to a computer or used for other purposes. It is important that the amplitude (independent of noise) of the zero reference signal and the signal whose phase is to be measured be

constant. This procedure is often used to monitor the relative phase change, as an ongoing function of time, of a signal that does not vary in amplitude.

(1) Measure the amplitude of the zero-phase reference signal and the signal whose phase is to be measured in the normal manner, and record the amplitudes measured. If the relative phase variations of a signal are to be measured, however, the amplitude need not be known but it must be constant.

(2) Apply the zero-phase reference signal to the input.

(3) Set up the Signal Channel filter parameters. If the frequency is constant, use a Q of 100 and the Bandpass mode. If the frequency is changing, it would be best to use a wide bandwidth and to avoid using a transformer input, because internal phase and amplitude variations as a function of frequency could not be accounted for.

(4) Set the Zero Offset toggle to neutral. Then adjust the Phase controls for a positive peak. Also, if the Bandpass mode is used, fine-adjust the frequency controls for a positive peak.

(5) Adjust the fine sensitivity (screwdriver) control for an exact + full scale meter indication. (This throws the gain calibration off, so after the phase measurement is complete, the instrument should be recalibrated.) The Sensitivity switch setting at which this full-scale adjustment is made is referred to below as the "reference full-scale range".

Because of the limited range of the fine sensitivity control, for a fixed reference amplitude it is not always possible to adjust for exact full-scale meter indication. In such a situation a different level of reference signal should be used. If this is not possible, an intermediate level on the scale can be referred to as "full-scale". However, the following procedure and readings must be modified accordingly.

(6) Set the Zero Offset control to "+", and turn the Offset vernier exactly ten turns clockwise from zero. The overload lamp will light, and the meter will peg downscale.

(7) Increase the sensitivity by a factor of 10 (sensitivity control 3 positions ccw), so that the meter again reads on scale.

(8) Adjust the Phase dial for an up-scale peak. If operating in the Bandpass mode, alternately adjust the Vernier and Signal Channel Frequency fine controls for an up-scale peak. Alternate between the two adjustments until no further increase in the meter indication can be obtained.

(9) Adjust the fine sensitivity screwdriver control for an exact meter null (which, incidentally, corresponds to a more exact full-scale setting for the 10 x less sensitive range).

(10) Change the Phase quadrant 90°, and return the Zero Offset toggle to neutral (center position). The meter should read near zero, but the small phase error will probably cause a reading slightly off null. Adjust the Phase dial control for exact meter null. Turn the Sensitivity switch counterclockwise as far as possible without overload while making this adjustment.

(11) Return the Sensitivity switch to the reference full-scale range setting.

(12) Apply the signal whose phase is to be measured to the Preamplifier input. The meter indication with respect to the unity meter scale is accurately equal to the cosine of the phase angle if the reference signal and measured signal are exactly equal in amplitude. Because the full-scale output is 10 V, the output voltage is 10 x the cosine of the phase angle. If the reference signal and the signal whose phase is being measured are unequal in amplitude, the cosine function must also be multiplied by the ratio of the amplitudes of the two signals, V_{ref}/V_X. The amplitudes were measured in step (1).

For small angles, much higher resolution can be obtained by increasing the sensitivity. Remember, however, to always refer voltages and meter readings obtained with increased sensitivity back to the full-scale reference range by a multiplier equal to the ratio of the ranges.

(13) Use a cosine table or a computer to convert the readings and voltages obtained to the phase angle.

3.10 REAR PANEL CONNECTORS

3.10A INTERFACE CONNECTOR (J9)

J9 is a 14-pin connector having outputs as given in Table III-7. This connector mates with Amphenol #57-30140, and is wired for compatibility with the Model 127, which is a two-phase accessory. If it is desired to operate a Model

Pin	Signal
1	Ground
2	+24 V dc
3	−24 V dc
4	No connection
5	0° Reference
6	90° Reference
7	180° Reference
8	Signal Out
9	Signal Ground
10	No connection
11	VCO Input
12	Reference Input
13	No connection
14	270° Reference

Table III-7. INTERFACE CONNECTOR SIGNALS AND PINS

123 AC Zero Offset Accessory with the Model 124A, a special cable is available which interconnects between the Model 123 and J9 of the Model 124A. Note that the VCO Input (pin 11) is also accessible by means of a rear-panel BNC jack.

3.10B EXT. TIME CONSTANT

J8 is the External Time Constant socket. By connecting external capacitors to the proper pins of this socket, time constants in excess of 300 s can be obtained. Two capacitors are required, one to be connected between pins 8 and 9, and the other between pins 10 and 11. The resulting time constant is $3C \times 10^7$ seconds, where C is the capacitance (single capacitor) in farads. A table of the signals provided at this connector follows.

Pin	Function
1	Ground
2	+24 V (maximum of 100 mA)
3	–24 V (maximum of 100 mA)
4	No connection
5	–31 V in (for battery operation)
6	No connection
7	+31 V in (for battery operation)
8	One lead of first time constant capacitor
9	Other lead of first time constant capacitor
10	One lead of second time constant capacitor
11	Other lead of second time constant capacitor

Table III-8. EXTERNAL TIME CONSTANT CONNECTOR SIGNALS AND PINS

3.11 BATTERY OPERATION

Battery operation of the Model 124A Lock-In Amplifier may be necessary where no ac power is available, or as a last resort where power line interference is a problem. Battery operation is particularly straightforward because the necessary internal points are available at the rear-panel 11-pin socket. Two batteries are required, one to supply +31 V (400 mA) and the other to supply –31 V (360 mA). The +31 V source should be connected to pin 7. The –31 V source should be connected to pin 5. Ground for both is at pin 1. It is generally a good idea to fuse the battery lines external to the instrument, and to provide an ON/OFF switch as well. The front-panel ON/OFF switch is not functional when the instrument is operated from batteries. The line cord should be disconnected. Other than the already mentioned ON/OFF switch not functioning, there is only one other point of difference between a battery operated instrument and one operated from the line, and that is that the pilot lamps which illuminate the panel meter will not light. Because of the ac power requirements of the digital panel meter, UNITS INCORPORATING THE DIGITAL PANEL METER OPTION CANNOT BE OPERATED FROM BATTERIES.

3.12 PHASE METER MODIFICATION (1241/85)

A Model 124A equipped with this option can be operated either as a normal lock-in amplifier or as a Phase Meter, depending on the position of the rear-panel NORMAL/PHASE switch.

When the instrument is operated as a phase meter, the input signal, after some initial ac gain, is routed through an amplifier limiter that has a constant amplitude (clipped) rectangular output. This signal, when synchronously demodulated, yields a dc output that is a linear function of the phase difference between the reference and input signals. The phase sensitivity is 100 mV out per degree with the Function switch set to LOW DRIFT. Only LOW DRIFT operation can be used and the Sensitivity switch is constrained to settings in the range of 1 μV to 500 mV. In the case of units equipped with the Digital Panel Meter option, Phase measurements can only be made with the Sensitivity switch set to 1 μV, 10 μV, 100 μV, 1 mV, 10 mV, or 100 mV. In other words the "2" and "5" positions should not be used for making phase measurements if the unit is equipped with a digital panel meter.

In phase-meter operation the input signal should be limited to less than ten times full scale (but not more than 200 mV) for Sensitivity switch settings from 1 μV to 100 mV. For the 200 mV and 500 mV sensitivity positions, the maximum input signal is 500 mV. The phase indication will not be in error by more than 5° maximum providing the signal amplitude is at least 100 μV or 20% of full scale as indicated by the setting of the Sensitivity switch, whichever is greater.

3.13 MIXER MONITOR MODIFICATION (1241/92)

Units equipped with the Mixer Monitor Option have an additional rear-panel BNC connector. The signal available at this output is taken directly from the output of the Mixer and before any filtering. Figure III-24 illustrates the Mixer output corresponding to in-phase and quadrature signals respectively. If the signal and reference inputs to the Mixer are either in phase or 90° out-of-phase, the signal at the output of the Mixer will be as shown. For signals 180° out-of-phase, the Mixer output will be the inverse of the 90° output. Taking the maximum possible area that can be enclosed by one cycle (one polarity) as a unit output, the output averaged over a cycle for any Mixer input phase relationship is simply the unit output times the cosine of the angle between the input and reference signals.

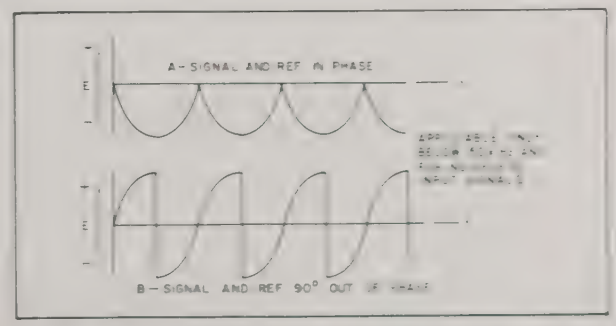


Figure III-24. MIXER OUTPUT FOR IN-PHASE AND QUADRATURE SIGNALS

The cosine response depends on the sinusoidal nature of the input signal. If the signal were a square wave and the tuned amplifier were not used, the Mixer output would vary linearly with the angle between the signal and reference inputs. Nevertheless, maximum output would still be at 0° and 180°, and zero output would be obtained at 90° and 270°. Note that when the Model 124A is being operated in the Phase Meter mode (assuming the instrument in question is equipped with that option), internal limiting circuitry "converts" any input to a rectangular wave of the same period, and it is this rectangular wave that would be observed at the Mixer Monitor connector.

The amplitude of the Mixer Monitor output is 555 mV peak with a full-scale input signal at 0° and operating in the LO DRIFT mode. Operated in NORMAL, the Mixer Monitor signal decreases to 55.5 mV for a full scale input and in HI, it decreases to 5.55 mV. The output resistance is 1000 ohms.

It might be mentioned that the waveforms illustrated in Figure III-24 apply only at frequencies below 50 kHz and with a noise-free input signal. At higher frequencies, switching spikes become visible and some Mixer filtering effects become evident. Even relatively small amounts of input noise can completely obscure the signal at the Mixer Monitor output, especially in FLAT mode operation.

3.14 REMOTE PROGRAMMING OPTION MODIFICATION (1241/83)

In units equipped with this option, the Sensitivity and Dynamic Range Tradeoff can be remotely controlled by applying logic 0 (ground) to the appropriate pins of the rear-panel Remote Interface connector, J8001. Associated with the connector is a pushbutton switch that transfers the instrument from local to remote operation and vice versa. In Local operation, the Sensitivity and Dynamic Range Tradeoff are controlled by the front-panel controls in the usual manner. In Remote operation, these parameters are independent of the front-panel control settings and are determined instead by the inputs to the Remote Interface connector. Table III-9 indicates the pin assignments of this connector. Note that there are three "groups" of control input lines. To obtain any given combination of Sensitivity and Dynamic Tradeoff, one input in each group is grounded. Usually, all of the other pins can be left "floating". However, in a noisy environment, particularly where the cable leading to J8001 is relatively long, it may be advisable to apply logic 1 (+3.5 V ±1 V) to the other active input pins to assure stable operation. Otherwise, transient pickup could cause undesired "switching" of the Sensitivity and Dynamic Range Tradeoff. A connector (AMPHENOL 57-30360) that mates with J8001 is supplied with the modification.

In addition to the three groups of input lines, two outputs, OVERLOAD and REF. UNLOCK, are provided. Each of these outputs is "up" when the corresponding lamp is illuminated, and "down" when the corresponding lamp is dark.

There are two operating "restrictions" that the operator should bear in mind when operating a unit equipped with this modification. First of all, the Dynamic Tradeoff Over-rides that occur as a function of selected (programmed) Sensitivity apply in Remote Programmed operation exactly the same as in Local operation. (For details, see page III-15.) Second, there is a reduction in the amount of ±24 V power available for external use. From Table III-8, ±100 mA are available. In the case of units equipped with the Remote Programming Option, these levels are reduced to ±80 mA.

Pin	Use
1	OVERLOAD - logic "0" = lamp off
2	REF. UNLOCK - logic "1" = lamp on
3-18	No connection
19	100 nV Sensitivity
20	1 μV Sensitivity
21	10 μV Sensitivity
22	100 μV Sensitivity
23	1 mV Sensitivity
24	10 mV Sensitivity
25	100 mV Sensitivity
26	LO DRIFT Tradeoff
27	NORMAL Tradeoff
28	HI DYNAMIC RANGE (Reserve) Tradeoff
29-32	No connection
33	X1.0 Sensitivity Multiplier
34	X5.0 Sensitivity Multiplier
35	X2.0 Sensitivity Multiplier
36	Ground

Table III-9. REMOTE PROGRAMMING CONNECTOR PIN ASSIGNMENTS

In the case of an instrument equipped with both the Remote Programming Option and the Digital Panel Meter option, there are some special considerations that must be observed for proper operation. In units equipped with the Digital Panel Meter option alone, switching controlled by the front-panel Sensitivity switch sets the digital display sensitivity. In the case of units equipped with the Remote Programming option, the Sensitivity switch is rendered ineffective when the sensitivity is being controlled remotely, and control over the digital display sensitivity is lost.

To solve this problem, an additional toggle switch has been added to the rear panel of instruments equipped with both options. This switch has two positions, NORM, and D.P.M. 1.000. For operation with the remote-programming option *inactive* (this option is controlled by a rear-panel pushbutton), the switch should be set to NORM, in which position the display functions exactly as described in Subsection 3.8. In remote-programming operation, the switch should be set to D.P.M. 1.000, in which position the digital panel meter indicates the input signal level as a fraction of *full scale*, independent of the selected sensitivity. A full-scale input gives a display indication of 1.000, independent of whether the programmed sensitivity is 1 mV, 2 mV, 5 mV, or some other value. Similarly, with a full-scale input applied, the BCD output will be 1.000 and

the recorder output will be 10 V. The display indication and output levels are proportionally less with less than full-scale inputs. For example, if the programmed sensitivity is 200 mV, and a 100 mV signal is applied, the digital display will indicate 0.500 (half scale), the BCD output will be 0.500, and the recorder output will be 5 V.

3.15 SELECTIVE EXTERNAL REFERENCE MODIFICATION (1241/77)

In some applications, it may happen that the reference signal produced by the experimental apparatus is of very poor quality, that is, it is accompanied by much noise and interference. As explained earlier in the manual, use of a simple low-pass filter in series with the reference signal will usually clean up such a signal sufficiently to make it acceptable to the Model 124A reference circuits. Nevertheless, there could arise situations where this relatively simple technique would prove inadequate. If this is the case, the best one can do is to pass the reference signal through a tuned bandpass filter of moderate Q, perhaps 10. Even the poorest reference waveform, once it has been passed through such a filter, will be of sufficiently good quality to allow normal reference channel operation. The sacrifice one makes in using such a filter is that, for all practical purposes, the tracking capability of the Reference Channel is given up. Any change in the frequency of the reference signal results in amplitude loss and phase shift as the frequency moves out of the center of the passband.

Such a filter can, of course, be connected externally. However, in the case of Model 124A's equipped with the Selective External Reference Modification, a Q-of-10 filter is provided internally. These units are equipped with a rear-panel switch that allows the Selective External (tuned) Reference mode to be selected. With the switch in the NORMAL position, the instrument works exactly as described previously. With the switch in the SEL. EXT. position, the instrument operates in the Selective External Reference mode *providing* the front-panel Reference Mode switch is set to INT/VCO. If the front-panel Reference Mode switch is in any other position when the rear-panel switch is set to SEL. EXT, improper Reference Channel operation results.

The only other consideration in using the Selective External Reference mode is to tune the Reference Channel to the frequency of the input reference signal. This is easily done by applying the reference signal, and then monitoring the amplitude of the signal at the Reference OUT connector with the Signal Channel. A suitable procedure follows.

(1) Set the controls as follows.

Power: ON
Selective External Selector switch (rear panel): SEL. EXT.
Reference Mode switch: INT/VCO
Reference Level switch: 10 (if preamp. is Model 118, set to 1)
Reference Level Vernier: CAL.
Function switch: ACVM
Sensitivity: 500 mV
Signal Mode: FLAT
Signal Input switch: A

(2) Set the Reference Frequency controls to the approximate frequency of the Reference signal.

(3) Connect the Reference signal to the REF IN connector. **NOTE:** The amplitude of the Reference signal should be in the range of 100 mV pk-pk to 3 V pk-pk.

(4) Connect a cable from the REF OUT connector to the "A" Input connector of the preamplifier. The panel meter should show some deflection. If it does not, adjust the Reference Frequency controls as required to obtain some deflection and then further adjust them for maximum meter indication. If the meter indication exceeds full scale, use the Reference Level vernier to reduce the indication to about 50% of full scale. Then readjust the Reference Frequency controls for the desired maximum indication.

This completes the tuning procedure. The internal bandpass filter is now tuned to the reference frequency. The cable interconnecting the REF OUT connector and the "A" Input of the Preamplifier can now be removed and the instrument operated in the usual manner, bearing in mind that, if the reference frequency is changed, retuning will be required.

SECTION IV SAFETY NOTICE

WARNING! POTENTIALLY LETHAL VOLTAGES ARE PRESENT INSIDE THIS APPARATUS. THESE SERVICE INSTRUCTIONS ARE FOR USE BY QUALIFIED PERSONNEL ONLY. TO AVOID ELECTRIC SHOCK DO NOT PERFORM ANY SERVICING OTHER THAN THAT CONTAINED IN THE OPERATING INSTRUCTIONS UNLESS YOU ARE QUALIFIED TO TO SO. Any adjustment, maintenance and repair of the opened apparatus under voltage shall be avoided as far as possible and, if unavoidable, shall be carried out only by a skilled person who is aware of the hazard involved. When the apparatus is connected to a power source, terminals may be live, and the opening of covers or removal of parts is likely to expose live parts. The apparatus shall be disconnected from all voltage sources before it is opened for any adjustment, maintenance, or repair. Note that capacitors inside the apparatus may still be charged even if the apparatus has been disconnected from all voltage sources. Service personnel are thus advised to wait several minutes after unplugging the instrument before assuming that all capacitors are discharged. If any fuses are replaced, be sure to replace them with fuses of the same current and voltage rating and of the same type. The use of makeshift fuses and the short-circuiting of fuse holders are prohibited.

SECTION IV
ALIGNMENT PROCEDURE

READ SAFETY NOTICE ON PRECEDING PAGE BEFORE PROCEEDING

4.1 INTRODUCTION

The Model 124A Lock-In Amplifier is a reliable conservatively designed instrument. High quality stable components have been used throughout in its construction and one can reasonably expect a long period of troublefree operation without any need for realignment. However, to be assured of continued high confidence in the experimental data obtained with the Model 124A, it may be advisable to run through the following alignment at one year intervals, and after doing a repair on the instrument. Due to possible interaction between some of the adjustments, it is necessary that they be carried out in the indicated sequence. Any decision to make a partial alignment should be reserved to someone having sufficient knowledge of the Model 124A to fully understand all possible interactions. Figure IV-1 identifies the adjustments and board-edge testpoints. To identify the gold pin-type testpoints by their "TP" number (these testpoints are not located at the board edge), it will be necessary to refer to the appropriate individual-board parts location diagram in Section VI. Some of these testpoints are also identified by an "E" or "B" number printed on the board. This number also appears in the text references, allowing the testpoints to be easily identified.

Note that this alignment procedure is not intended to be used in troubleshooting. If the unit is suspected to be malfunctioning, go directly to Section V, which deals with troubleshooting. The instrument must be working normally before it can be aligned.

4.2 EQUIPMENT NEEDED

(1) DC Voltmeter with "center zero". A dc coupled scope may be used instead.

(2) Digital Voltmeter.

(3) General purpose oscilloscope.

(4) General purpose sinewave generator.

(5) Frequency counter.

(6) AC Voltmeter such as the HP Model 400EL.

(7) Two BNC shorting plugs, CW-159/U (Amphenol or equivalent).

(8) Extender Board, Princeton Applied Research #1710-00-1403S.

(9) Nonmetallic alignment tool to be used for high frequency "screwdriver adjustments".

4.3 PROCEDURE

4.3A PRELIMINARY STEPS

(1) Plug in any of the following preamplifiers: Model 116 operated direct, Model 117, Model 118, or Model 119 operated direct. **NOTE:** If a Model 118 or Model 185 Preamplifier is used, it will be necessary to take into account the factor of ten higher gain of this preamplifier. This is done by always selecting a sensitivity that is a factor of ten lower than that called for in the procedure.

(2) Connect the BNC shorting plugs to both inputs of the Preamplifier.

(3) Remove the top and bottom covers. The top cover slides off to the rear after removing the two screws underneath the upper cover overhang at the rear of the instrument. The bottom cover slides off to the rear after removing the two screws which secure the two rear bumper feet.

(4) Set the Model 124A controls as follows.

Sensitivity: 500 μV
Mode: FLAT
Signal Frequency Digits: 4.00
Signal Frequency Multiplier: X100
Signal Q: 1
Reference Frequency Controls: NORM, 4.00, X100
Reference Mode: INT/VCO
Reference Level: 10
Reference Level Vernier: CAL (fully clockwise)
Phase dial: 0.00°
Phase switch: 0°
Time Constant: MIN.
6 dB/12 dB switch: 6 dB
Zero Suppress Toggle switch: OFF (center position)
Zero Suppress dial: 0.00
Function: HI DYNAMIC RANGE
Calibrate: 100 mV
Power: ON

(5) Allow a fifteen minute warmup.

4.3B ±24 V ADJUSTMENTS (R6028 and R6010) - POWER SUPPLY BOARD

(1) Monitor the voltage at TP6002 (yellow testpoint) with the digital voltmeter (referred to hereafter as DVM).

(2) Adjust R6028 (–24 V ADJ) for a DVM indication of –24.0 V.

(3) Transfer the DVM to TP6000 (red testpoint).

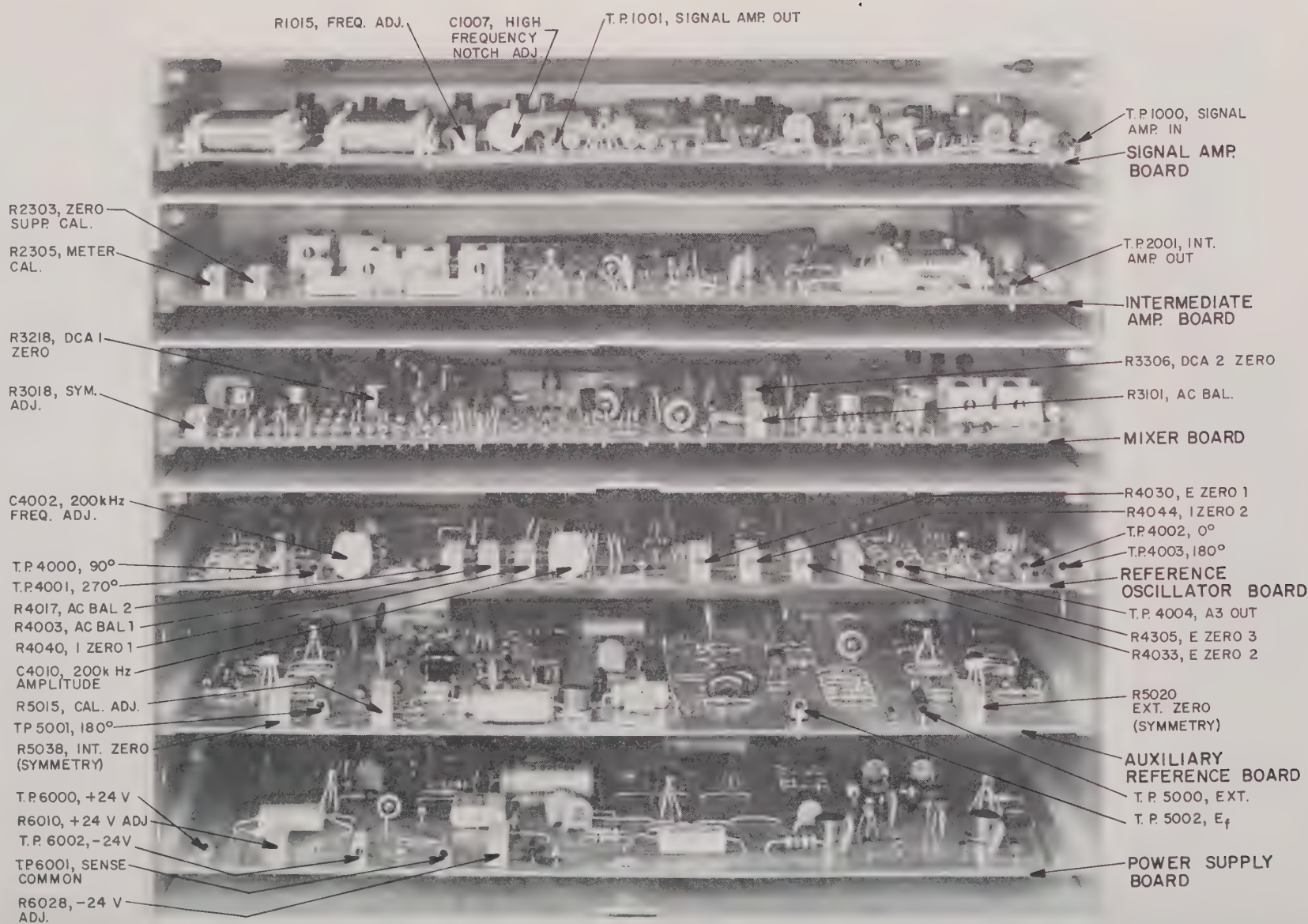


Figure IV-1. MODEL 124A ADJUSTMENTS AND TESTPOINTS

(4) Adjust R6010 (+24 V ADJ) for a DVM indication of +24.0 V.

4.3C INITIAL REFERENCE OSCILLATOR BOARD ADJUSTMENTS

(1) Turn off the power. Then remove the Reference Oscillator board and plug the Extender board into the unit in place of the Reference Oscillator board.

(2) Plug the Reference Oscillator board into the Extender board, turn on the power, and allow a five-minute warmup.

(3) E ZERO 3 ADJUST (R4305)

(a) Connect the dc voltmeter (not the DVM) to TP4004 (gray testpoint).

(b) Connect a jumper between CR4007 and R4050 as indicated on the Parts Location Diagram on page VI-16. Under no circumstances use chassis ground as the circuit may oscillate.

(c) Adjust R4305 (E ZERO 3 ADJ) for 0 V ±1 V at the testpoint. Note that this is an "open loop" high gain adjustment, and so will be difficult to set, and, once set, will drift quickly from the ideal "0" reading.

(d) Remove the jumper. The dc voltage at the gray testpoint should stabilize at −3.8 V ±0.5 V.

(4) I ZERO 1 ADJUST (R4040)

(a) Connect two jumpers, one from TP4005 (B1) to ground and the other from TP4006 (E1) to ground. TP4005 and TP4006 are both gold-pin testpoints down on the board.

(b) Connect the voltmeter to TP4001 (green testpoint).

(c) Adjust R4040 (I ZERO 1 ADJ) so that the monitored voltage is drifting equally about zero.

(d) Remove the jumper which extends from TP4006 (E1) and ground, but leave the jumper which extends from TP4005 (B1) and ground.

(5) E ZERO 1 ADJUST (R4030)

(a) Adjust R4030 (E ZERO 1 ADJ) for equal drift about zero in the monitored voltage (voltmeter still connected to TP4001).

(b) Remove the jumper which extends from TP4005 (B1) and ground.

(6) I ZERO 2 ADJUST (R4044)

(a) Connect two jumpers, one from TP4007 (B2) and ground, and the other from TP4008 (E2) and ground. These are both gold-pin testpoints down on the board.

(b) Connect the voltmeter to TP4003 (violet testpoint).

(c) Adjust R4044 (I ZERO 2) such that the monitored voltage drifts equally plus and minus about zero.

(d) Remove the jumper which extends from TP4008 (E2) and ground, but leave the jumper which extends from TP4007 (B2) and ground.

(7) E ZERO 2 ADJUST

(a) Adjust R4033 (E ZERO 2 ADJ) for equal drift about zero as measured at TP4003 (violet testpoint).

(b) Remove the jumper which extends from TP4007 (B2) and ground.

(8) Turn off the power. Then remove the Reference Oscillator board from the extender, remove the extender, and return the Reference Oscillator board to its proper place in the instrument. Turn the power back on.

4.3D AUXILIARY REFERENCE BOARD ADJUSTMENTS

(1) INTERNAL ZERO SYMMETRY ADJUST (R5038)

(a) Connect the oscilloscope to the front-panel CALIBRATE OUT connector. The sweep time should be 0.2 ms/cm and the oscilloscope should be adjusted to trigger on the positive slope of a 222 mV pk-pk square wave.

(b) Carefully note the duration of the positive half cycle of the square wave. Then trigger on the negative slope of the square wave and carefully note the duration of the negative half cycle. The two "half cycles" should have exactly the same duration. If they do not, adjust R5038 (INT ZERO ADJ) as required to obtain the desired symmetry.

(3) DC CALIBRATE ADJUST (R5015)

(a) Set the front-panel Calibrate switch to 222 mV dc.

(b) Remove the oscilloscope from the Calibrate jack and connect the DVM there instead.

(c) Adjust R5015 (DC CAL ADJ) for a DVM indication of −.2220 V. Remove the DVM.

(4) EXT. ZERO SYMMETRY ADJUST (R5020)

(a) Set the front panel Reference Level switch to ".2".

(b) Connect the oscilloscope to TP5000 (green test-point).

(c) Connect a cable from the REF. OUT jack to the REF. IN jack.

(d) While observing the square wave at TP5000, gradually rotate the front-panel Reference Vernier counterclockwise. As the control is adjusted, a point will be reached where the symmetry of the square wave will begin to degrade. When this occurs, adjust R5020 (EXT. ZERO SYMMETRY ADJ) as required to maintain as near perfect symmetry as possible. Continue until the waveform "locks" to either "+" or ground, indicating that the vernier is too far counterclockwise.

(e) Set the Reference Level switch to 10 and rotate the vernier to the fully clockwise (CAL) position.

(f) Remove the cable interconnecting the REF IN and REF OUT connectors.

4.3E MIXER BOARD ADJUSTMENTS

(1) Turn off the power. Then remove the Mixer board and plug in the Extender board in its place.

(2) Plug the Mixer board into the Extender board, turn the power back on, and allow a five minute warmup. Set the Function switch to LO DRIFT.

(3) DC AMP 2 ZERO ADJUST (R3306)

(a) Connect jumpers from TP3301 to TP3302 and to TP3303. Note that these are not board-edge testpoints but rather gold-pin testpoints down on the board.

(b) Adjust R3306 (DC AMP 2 ZERO) for "0" on the front-panel meter. This is a drifty open-loop adjustment.

(c) Remove the jumpers.

(d) Turn off the power, separate the Mixer board and the Extender, and remove the Extender. Then return the Mixer board to the instrument and turn the power back on.

(4) DC AMP 1 ZERO (R3218)

(a) Set the Function switch to HI DYN RANGE.

(b) Connect the DVM to the FUNCTION OUT jack.

(c) Adjust R3218 (DC AMP 1 ZERO) for 0.00 V at the DVM. Remove the DVM.

(5) AC BAL ADJ (R3101)

(a) Connect the oscilloscope to the front panel FUNCTION OUT connector.

(b) Set the Time Constant switch to MIN.

(c) Adjust R3101 (AC BAL ADJ) for minimum ripple as observed at the oscilloscope.

(d) Increase the time constant to 300 ms and remove the oscilloscope.

(6) SYMMETRY ADJUST (R3018)

(a) Change the instrument control settings as follows.

Signal Frequency Digits: 8.00
Signal Mode: BANDPASS
Signal Q: 20
Function: ACVM
Sensitivity: 10 mV
Reference Frequency controls: NORM, 3.(10) 1, X100 (net freq. = 401 Hz)

(b) Remove the shorting plug from the A Input of the Preamplifier. Then connect the External Signal Generator to the "A" Input. Be sure the Input Selector of the Preamplifier is set to "A". The frequency of the signal generator output should be 800 Hz.

(c) Adjust the amplitude of the signal generator output for an "on scale" indication on the Model 124A panel meter. Then carefully vary the frequency of the signal generator output for maximum deflection of the panel meter.

(d) Readjust the amplitude of the signal generator output for exactly full scale deflection of the Model 124A panel meter.

(e) Set the function switch to HI DYNAMIC RANGE and the Time Constant switch to 1 SEC.

(f) Set the Sensitivity switch to 100 μV.

(g) Adjust the Reference Frequency controls for a panel meter "beat" of about 1 Hz.

(h) Adjust R3018 (SYMMETRY ADJ) for minimum pk-pk amplitude in the observed beat.

(i) Set the Phase switch to 90°. Then note and record the pk-pk amplitude of the beat.

(j) Adjust R3018 (SYMMETRY ADJ) as required to reduce the amplitude of the "beat" by exactly one half.

(k) Reset the Phase switch to 0° and check to see

that the "beat" is the same amplitude on 0° as it is on 90°.

(l) Reset the Time Constant to 300 ms. Remove the signal from the input and reconnect the shorting plug removed in step b.

4.3F INTERMEDIATE AMPLIFIER BOARD ADJUSTMENTS

(1) ZERO SUPPRESS CAL (R2303)

(a) Set the Function switch to LO DRIFT.

(b) Set the Zero Suppress Polarity switch to "–".

(c) Set the Zero Suppress dial to 1.00 (one turn from the fully counterclockwise position).

(d) Connect the DVM to the front-panel FUNCTION OUT connector. Then adjust R2303 (ZERO SUPPRESS CAL) for a DVM indication of +10.00 V.

(2) METER CAL (R2305)

(a) Note the panel meter indication. It should be near full scale to the right.

(b) Adjust R2305 (METER CAL) for exactly full-scale panel meter deflection.

(c) Set the Zero Suppress Polarity switch to the center (OFF) position.

4.3G FINAL REFERENCE OSCILLATOR BOARD ADJUSTMENTS

(1) AC BAL 1 and AC BAL 2 ADJUSTMENTS (R4003 and R4017)

(a) Connect the ac voltmeter to TP4000 (white testpoint).

(b) Set the Model 124A Reference Frequency controls to NORM, 4.00, X100 (400 Hz).

(c) Set the Reference Level switch to "1" and the Reference Vernier fully clockwise.

(d) Connect the Frequency Counter to the REF. OUT connector.

(e) Carefully note the signal level at TP4000. Then transfer the ac voltmeter to TP4002 (blue testpoint) and note the signal level there.

(f) Adjust R4003 (AC BAL 1) so that the amplitude at TP4002 is the same as it is at TP4000.

(g) Note the *difference* between the frequency indicated by the counter (about 400 Hz) and the frequency set by the Reference Frequency controls (exactly 400 Hz). Then readjust R4003 (AC BAL 1 as required to reduce the *difference frequency* by exactly a factor of two.

(h) Adjust R4017 (AC BAL 2) for a counter indication of exactly 400 Hz.

(2) HIGH FREQUENCY AMPLITUDE ADJUST (C4010) and 200 kHz FREQ. ADJ (C4002)

(a) Set the Reference Frequency controls to NORM, 2.00, and X10K.

(b) Note and record the frequency indicated on the counter.

(c) Set the Reference Frequency controls to ADD 10, 10.00, and X10K.

(d) Connect the ac voltmeter to TP4002 (blue testpoint).

(e) Adjust C4010 (HIGH FREQ. AMP. ADJ) for an ac voltmeter indication of exactly 1 V rms. Be sure to use the non-metallic alignment tool for this adjustment.

(f) Adjust C4002 (200 kHz FREQ. ADJ) for a counter indication exactly ten times that noted in step "b".

(g) Reset the Reference Frequency controls to NORM, 2.00, and X10K. The frequency should be the same as noted in "b". If it has changed, record the new frequency, and then set the Frequency controls back to ADD 10, 10.00, and X10K. Readjust C4002 (200 kHz FREQ. ADJ) to obtain a counter reading exactly ten times the new frequency.

4.3H SIGNAL BOARD ADJUSTMENTS

(1) SIGNAL FREQUENCY ADJUST (R1015)

(a) Set the controls as follows.

Signal Frequency controls: 4.05 X100 (405 Hz)
Reference Frequency controls: NORM, 4.05 X100 (405 Hz) (**Note:** Do not change Reference controls during this check)
Signal Mode: BANDPASS
Function: ACVM
Signal Q: 100
Calibrate: 2 mV
Sensitivity: 2 mV
Preamp Input Selector: A

(b) Remove the shorting plug from the "A" Input of the preamplifier and connect a cable from the "A" Input to the Calibrate Output.

(c) Adjust the third dial of Signal Frequency control for peak panel meter indication.

(d) Change the setting of the Signal Q switch from 100 to 10% ENBW. The signal amplitude should not change. If it does, change the setting of the front-panel NOTCH ADJUST screwdriver adjustment as required so that no amplitude change takes place when the Q is switched from 100 to 10% ENBW. Leave the Q set to 100.

(e) Now set the *third* Signal Frequency dial to "5". (**Note**: Both Signal and Reference Frequency controls should be set the same.)

(f) Adjust R1015 (SIG. FREQ. ADJ) for maximum panel meter indication.

(2) HIGH FREQUENCY NULL ADJUSTMENT (C1007)

(a) Set the controls as follows.

Signal Frequency controls: 10.95, X10K (109.5 kHz)
Reference Frequency controls: NORM, 10.95, X10K (109.5 kHz)

(b) Adjust the third dial of the Reference Frequency controls for peak panel meter indication. Carefully note the panel meter indication.

(c) Set the Q Selector to 10% ENBW. If the meter indication changes, adjust C1007 (HIGH FREQ. NULL) so that there is no meter indication change as the Q is switched back and forth between 100 and 10% ENBW. Leave the Q set to 10% ENBW.

(d) Disconnect the cable which extends from CAL OUT to the A INPUT.

4.3I FINAL ADJUSTMENTS

The following adjustments can be made only after the instrument has been thoroughly warmed up with the cover in place. At the factory, a special top cover is used, one having holes drilled in it to give access to R3101, R3218, and C1007. The first two of these adjustments are located on the Mixer board. The third is located on the Signal board. It is not expected that the person doing the alignment will drill holes in his cover. By substituting a piece of cardboard for the cover the alignment can be successfully completed. Be sure the holes are accurately located and no larger than they have to be. With the "cover" in place, allow a one hour warmup before proceeding.

(1) AC BAL. ADJ (R3101) - MIXER BOARD

(a) Set the front-panel controls as follows.

Reference Frequency controls: NORM, 4.00, X10
Sensitivity switch: 5 mV
Time Constant: MIN.
Function switch: HIGH DYNAMIC RANGE

(b) Connect a shorting plug to the "A" Input.

(c) Connect the oscilloscope to the FUNCTION OUT connector.

(d) Adjust R3101 (AC BAL ADJ) so that the square wave ripple observed is minimum. It should be possible to get it below 400 mV pk-pk.

(2) HIGH FREQUENCY NULL ADJUSTMENT (C1007) - SIGNAL BOARD

(a) Set the controls as follows.

Signal Frequency: 10.95, X10K (109.5 kHz)
Reference Frequency: 10.95, X10K (109.5 kHz)
Function: ACVM
Calibrate and Sensitivity: 2 mV

(b) Remove the shorting plug from the "A" Input of the Preamplifier and connect a cable from the "A" Input to the Calibrate Output.

(c) Adjust the third dial of the Reference Frequency controls for peak panel meter indication. Carefully note the meter indication.

(d) Set the Q Selector to 10% ENBW. If the meter indication changes, adjust C1007 (HIGH FREQ. NULL) so that there is no meter indication change as the Q is switched back and forth between 100 and 10% ENBW.

(e) Remove the cable interconnecting the CAL output and "A" Input. Then return the shorting plug to the "A" Input.

(3) DC AMP 1 ZERO ADJ (R3218) - MIXER BOARD

(a) Set the controls as follows.

Sensitivity: 500 μV
Signal Frequency Digits: 4.00
Signal Frequency Multiplier: X100
Reference Frequency controls: NORM, 4.00, X100
Function: HIGH DYNAMIC RANGE

(b) Connect the DVM to the FUNCTION OUT connector.

(c) Adjust R3218 (DC AMP 1 ZERO ADJ) for 0.00 V at the DVM.

This completes the alignment. All test equipment can be removed and the covers secured in place.

4.3J PHASE METER OPTION ALIGNMENT

The following alignment is carried out only on units equipped with the Phase Meter Option. This procedure is to be performed *after* the regular alignment is completed.

(1) Remove the top cover. Then turn on the power and allow a fifteen minute warmup.

(2) Set the Model 124A controls as follows.

Input Selector (Preamplifier): A
Sensitivity: 50 mV
Signal Mode: FLAT
Reference Frequency controls: NORM, 4.00, X100
Reference Mode: INT
Reference Level switch: .1 V
Reference Level vernier: fully clockwise
Phase dial: 9.00 (90°)
Phase switch: 270°
Time Constant: 300 ms (6 dB)
Zero Offset dial: 0.00 (fully counterclockwise)
Zero Offset switch: OFF (center position)
Function switch: ACVM
NORM/PHASE switch (rear panel): PHASE

(3) Connect a cable from the Reference Channel OUT connector to the Preamplifier "A" Input.

(4) Connect the oscilloscope, dc coupled, to pin 9 of the Phase board. **NOTE:** This board is mounted against the Signal Amplifier board shield.

(5) The signal observed at the oscilloscope should be a 400 Hz square wave. After verifying that the signal is as indicated, rotate the front-panel Reference Level vernier counterclockwise as required to make the square wave noisy.

(6) Adjust the A2 DC control (located on the Phase board) for best symmetry in the observed square wave. When best symmetry is obtained, try further reducing the Reference Output amplitude and repeat the symmetry adjustment. Repeat these two steps until no further improvement in the setting of the symmetry adjustment can be made.

(7) Before proceeding, check at pin 5 of the Phase board for a sinewave output with an amplitude of approximately 20-to-60 mV pk-pk. Then set the rear-panel NORM/PHASE switch to NORM, and adjust the Reference Level Vernier for 1 V pk-pk at the oscilloscope.

(8) Connect the oscilloscope to J3-9 (Mixer board). Then set the Sensitivity switch to 5 mV and adjust the Reference vernier for 1 V pk-pk amplitude in the observed signal.

(9) Set the Sensitivity switch to 50 mV.

(10) Adjust the A1 GAIN trim-potentiometer (located on the Phase board) for 0.1 V pk-pk amplitude in the observed signal.

(11) Set the NORM/PHASE switch back to PHASE. Then set the Model 124A Function switch to LOW DRIFT.

(12) Observing the oscilloscope, adjust the front-panel Reference Level vernier until the amplitude of the monitored sinewave is just high enough to cause both negative and positive clipping.

(13) Adjust the CLIPPING SYMMETRY trim-potentiometer, located on the Phase board, for symmetrical clipping of the observed signal. Then remove the oscilloscope.

(14) Connect the DVM (digital voltmeter) to the front-panel Function OUT connector. Then adjust the A2 AMPLITUDE trim-potentiometer (located on the Phase board) for a DVM indication of 9.00 V.

(15) Carefully set the front-panel Phase dial for peak DVM indication. Then readjust the A2 AMPLITUDE trim-potentiometer for the desired 9.00 V reading.

This completes the Phase board alignment.

4.4 MODEL 116, 117, OR 119 PREAMPLIFIER ALIGNMENT

To align the preamplifier, it will be necessary to use a Model 183 Remote Preamplifier Adapter with extender cable.

4.4A PRELIMINARY STEPS

(1) Plug the Model 183 Remote Preamplifier Adapter into the Model 124A. Then interconnect the Model 183 and the Preamplifier with the Extender Cable.

(2) Remove the two screws at the rear of the preamplifier that secure the cover. Then slide the cover back onto the cable to get it out of the way.

(3) Set the Preamplifier Input Selector to "A–B". In the case of a Model 116 or a Model 119, set the Mode selector to DIRECT.

(4) Set the Model 124A controls as follows.

Sensitivity: 500 μV
Mode: FLAT
Signal Channel Frequency
  dials: setting immaterial
  multiplier switch: setting immaterial
Q Selector: setting immaterial
Reference Channel Frequency
  dials: 4.00, NORMAL
  multiplier switch: X100
Mode: INT/VCO
Phase switch: 0°
Phase dial: 90.0°
Reference Level: 1.0 V rms
Function: NORMAL
Calibrator Output Level: setting immaterial
Power: ON

4.4B PROCEDURE

(1) BIAS ADJUST (R101)

(a) Connect the DVM to that end of R130 which is in common with R127 and R133.

(b) Adjust R101 (BIAS ADJUST) for a voltage indication of –5.1 V.

(2) DC ZERO (R119)

(a) Transfer the DVM to either end of R132.

(b) Adjust R119 (DC ZERO) for a voltage indication of 0.00 V.

(3) COMMON-MODE REJECTION (R136)

(a) Connect the M124A Reference Channel output to *both* the A and B inputs of the preamplifier. **NOTE:** If an overload indication occurs, ignore it.

(b) Monitor the signal at R132 with an ac voltmeter or sensitive oscilloscope. Then adjust R136 (COMMON-MODE REJECTION) for a null in the ac voltmeter indication.

(4) HIGH-FREQUENCY COMMON-MODE REJECTION (C109)

(a) Change the setting of the Reference Frequency controls to 10.00, NORMAL, x10k. The frequency of the applied signal will now be 100 kHz. Set the M124A Reference output level to 0.5 V.

(b) While continuing to monitor the signal at R132 with the ac voltmeter (or oscilloscope), adjust C109 for a null in the ac signal level.

(5) DC ZERO CHECK

(a) Disconnect the calibrator output from the preamplifier inputs. Also, remove the ac voltmeter from R132 and connect the DVM to this point instead.

(b) The indicated voltage should be 0 V ±50 mV. If it is not, touch up the setting of R119 as required to obtain an indicated voltage of 0.00 V.

This completes the preamplifier alignment. The preamplifier cover can now be returned to its normal position.

4.5 MODEL 118 PREAMPLIFIER ALIGNMENT

To align the preamplifier, it will be necessary to use a Model 183 Remote Preamplifier Adapter with extender cable.

4.5A PRELIMINARY STEPS

Perform steps 1 through 4 inclusive of the procedure outlined in Subsection 4.4A.

4.5B PROCEDURE

(1) DC ZERO (R105 and R133)

(a) Connect the DVM to R125.

(b) Adjust R105 (DC ZERO) for an indicated voltage of 0.0 V.

(c) Alternate the voltmeter between resistors R125 and R144, and adjust R133 (DC ZERO) until the voltage at both points is the same.

(d) Readjust R105 (DC ZERO) for 0.0 V at R125.

(2) COMMON-MODE REJECTION (R130) AND RC BAL (R103)

(a) Set the Sensitivity to 1 mV. Then connect the M124A Reference output to *both* the A and B preamplifier inputs. Reduce the frequency to 40 Hz.

(b) Monitor the signal at R125 with the ac voltmeter. Then alternately adjust R130 (CMR ADJ.) and R103 (RC BAL) for a null in the measured signal level. Continue until no further improvement in the null can be obtained.

This completes the preamplifier alignment.

SECTION V
TROUBLESHOOTING

SECTION V SAFETY NOTICE

WARNING! POTENTIALLY LETHAL VOLTAGES ARE PRESENT INSIDE THIS APPARATUS. THESE SERVICE INSTRUCTIONS ARE FOR USE BY QUALIFIED PERSONNEL ONLY. TO AVOID ELECTRIC SHOCK, DO NOT PERFORM ANY SERVICING OTHER THAN THAT CONTAINED IN THE OPERATING INSTRUCTIONS UNLESS YOU ARE QUALIFIED TO TO SO. Any adjustment, maintenance and repair of the opened apparatus under voltage shall be avoided as far as possible and, if unavoidable, shall be carried out only by a skilled person who is aware of the hazard involved. When the apparatus is connected to a power source, terminals may be live, and the opening of covers or removal of parts is likely to expose live parts. The apparatus shall be disconnected from all voltage sources before it is opened for any adjustment, maintenance, or repair. Note that capacitors inside the apparatus may still be charged even if the apparatus has been disconnected from all voltage sources. Service personnel are thus advised to wait several minutes after unplugging the instrument before assuming that all capacitors are discharged. If any fuses are replaced, be sure to replace them with fuses of the same current and voltage rating and of the same type. The use of makeshift fuses and the short-circuiting of fuse holders are prohibited.

5.1 INTRODUCTION

This section consists of a series of procedures to be followed in troubleshooting the Model 124A. The purpose of the procedure is to narrow the trouble down to a specific circuit board by making voltage and waveform checks at critical points. Once the faulty board has been identified, the operator can contact the factory or one of its authorized representatives for advice on how to get the instrument back into operation in the shortest possible time. It may prove expedient to simply exchange the board for a new one. In the case of units still in Warranty, it is particularly important that the factory or one of its authorized representatives be contacted before doing any repair work on the board itself, because any damage that occurs as a result of unauthorized work could invalidate the Warranty.

Although past experience indicates that most instrument failures turn out to be the fault of a specific component failure on one of the boards, it is of course perfectly possible that some component other than the one located on a circuit board could go bad. Where this is the case, the person troubleshooting will have to appropriately adapt the procedure to isolate the faulty component.

In general, it is suggested that the person who carries out the troubleshooting procedure be well grounded in basic transistor electronics. The procedure is more to be thought of as a general guide for an experienced repairman than as a minutely detailed treatise to educate the newcomer.

5.2 EQUIPMENT REQUIRED

(1) General purpose oscilloscope.

(2) DC Voltmeter.

(3) Signal Generator able to supply a 1 V rms sinewave at 1 kHz.

(4) Extender Board, Princeton Applied Research #1710-00-1403S. This item is not really required for any of the checks called for in the following pages. However, it will prove indispensible for the troubleshooter who wants to go a little beyond the checks provided to isolate the trouble more specifically than is possible with the procedure. In using the board, be sure to install and remove circuit boards with the power off.

5.3 INITIAL STEPS

(1) Remove the top cover. It slides off to the rear after the two screws which secure it are removed. These two screws are located on the underside of the upper cover overhang at the rear of the instrument.

(2) After removing the hold down strap, lift each circuit board and give each a brief visual damage inspection. If any "charred" or otherwise damaged components are noticed, there is little point in going further.

(3) Be sure to check the fuses. There are two on the Power Supply board and one at the rear panel. They are discussed in Subsection 5.4, which follows.

5.4 POWER SUPPLY

(1) On the Power Supply board, check the voltage at TP6000 (red testpoint) for +24 V and at TP6002 (yellow testpoint) for −24 V. If the voltages are correct, go on to Subsection 5.5. If the voltages are incorrect or missing, proper power supply operation must be established before any further checks can be made. Note from the schematic on page VI-23 that the −24 V regulator supplies the reference voltage for the +24 V regulator. Thus, any trouble with the 24 V supply would cause loss of regulation in the +24 V circuit as well.

(2) Note that the unregulated input to both regulators is fused. If a check shows one of these fuses to be blown, try replacing it once. If it blows again, it will be necessary to locate and repair the short. One way to narrow the short down is to pull all boards but the Power Supply board (power off when boards are removed or replaced), and then to turn the power back on. If the fuse still blows, the trouble is most likely on the Power Supply board. If it does not blow, the board having the short can be easily determined by returning them one at a time until the fuse blows.

(3) If the regulator input fuses are not blown, but the +24 V levels are missing or incorrect, check the unregulated supply levels (nominally ±31 V) to isolate the problem to the Power Supply board or to the Unregulated Supply components (line fuse, transformer, rectifiers, or filter capacitors). Note that the high-power transistors are not located on the Power Supply board, but are instead mounted on the same plate as the transformer and filter capacitors.

5.5 REFERENCE CHECKS

(1) Control Settings

Meter: Check the mechanical zero and adjust it if necessary.

Preamplifier: Plug in a Model 117 Preamplifier, or a Model 116 or 119 operated in the DIRECT mode. A Model 118 or Model 185 can also be used, but certain Signal Channel checks will have to be modified as indicated in the text.

Sensitivity: 1 mV
Signal Channel Mode: FLAT
Reference Channel Mode: INT/VCO
Reference Frequency: 3.99, X100 (399 Hz)
Reference Level: 1 V rms (red inner knob fully clockwise)
Phase switch: 0°
Phase dial: 90°
Time Constant: 300 ms, 6 dB/octave
Zero Offset
dial: 0.00 (fully counterclockwise)
toggle switch: center (OFF) position
Function switch: NORMAL
Calibrator: 1 mV (100 μV if preamplifier is a Model 118)
Power switch: ON

(2) Reference Oscillator Board

(a) Check for −3.8 V ±05 V at TP4004 (gray testpoint on Reference Oscillator board). As indicated by the schematic on page VI-15, TP4004 monitors the output of A3, the Buffer between the front-panel Reference Frequency dials and the Voltage Controlled oscillator. If the voltage reading is correct, one can reasonably assume that the dial-controlled voltage dividers and Buffer A3 are functioning normally.

(b) Monitor TP4002 (blue testpoint) with the oscilloscope and check for a 2.8 V ±.3 V pk-pk sinewave at 399 Hz. **NOTE:** A faulty circuit will usually give indication of a "gross" error. For this reason it is not generally advisable to spend much time trying to determine whether the frequency or amplitude are "exactly" as specified. This applies to both this step and to the remainder of the procedure.

(c) Similarly check for a 2.8 V ±.3 V pk-pk at 399 Hz at TP4003 (violet), at TP4000 (white), and at TP4001 (green). These testpoints, together with TP4002, give access to the four outputs of the Reference Oscillator. If these signals are normal, one can be reasonably confident that the Reference Oscillator board is functioning normally.

(d) Transfer the oscilloscope to the Reference Output connector (front panel) and check for a 2.8 V ±.3 V pk-pk sinewave at 399 Hz. If this signal is as indicated, one can assume that the Reference Output Power Amplifier, located on the Power Supply board, is functioning normally.

(3) Auxiliary Reference Board

(a) Connect a 1 V pk-pk sinewave at 1 kHz from the signal generator to the Reference Channel IN connector.

(b) Set the Reference Channel Mode switch to EXT. The REF UNLOCK light should glow for a few seconds and then extinguish.

(c) Monitor the signal at the Reference Out connector with the oscilloscope. One should observe a 1 V rms (2.8 V ±.3 V pk-pk) sinewave at 1 kHz.

(d) Set the toggle switch to EXT f/2. Again, the REF UNLOCK light should glow, this time for about seven seconds, and then extinguish.

(e) Verify that the signal at the Reference Out connector is unchanged in amplitude but that its frequency has doubled (2 kHz).

(f) Reset the Reference Mode switch to INT/VCO but leave the signal generator connected to the Reference Input connector.

NOTE: If normal indications were obtained in steps "a" through "f", one can reasonably assume that the Auxiliary Reference board is functioning normally and go on to Subsection 5.6. If abnormal indications were noted, there is a good possibility of a malfunction on this board. The remaining steps in this sequence may prove helpful in narrowing the problem down to the specific malfunctioning circuit.

(g) +5 V Regulator: Check the voltage at the positive end of capacitor C5004. This voltage, which is

indicative of the current flow through Q5001 (and hence the +5 V load), should be +11 V ±3 V. C5004 is the 35 μF capacitor located near the upper edge of the board. Next check the +5 V regulator output, which should be +5 V ±0.5 V. This is most easily checked at the positive end of capacitor C5005, the 15 μF 6 V capacitor located near the upper edge at about the center of the board.

(h) Schmitt Triggers: Check for a 399 Hz square wave at TP5001 (blue testpoint). The upper and lower levels of the square wave should be +4.5 V ±1 V and 0.3 V ±0.3 V respectively. If the signal is as described, the Schmitt Trigger driven from the 180° output of the Reference Oscillator is functioning normally.

Next transfer the oscilloscope to TP5000 (green testpoint). A 1 kHz square wave with the same level limits as described in the preceding paragraph should be observed. If the signal is normal, one can conclude that the Schmitt Trigger driven from the External Reference signal is functioning normally.

(i) Frequency Comparator: A thorough checkout procedure for this circuit is beyond the scope of this manual. Nevertheless, failure to pass the following two tests is a clear indication of malfunction. If these tests are passed, no clear conclusions concerning the normality of these circuits can be made.

Test 1: Check for +4 V ±1 V at pin 11 of integrated circuit U5006. **NOTE**: IC packages U5001 through U5006 are labeled as "1" through "6" with labels consisting of small etched foil digits on the side of the board opposite the components. Pin 1 of each 14 digit package is similarly marked. The pins are counted clockwise (viewed from the label side).

Test 2: Check for a 399 Hz square wave at TP5002 (gray testpoint). The lower level of the square wave should be 0 V. There should be two upper levels, both between +3 and +4 V.

The signal generator should be disconnected from the Reference Input at this time.

NOTE: If steps "g" through "i" fail to identify the faulty circuit, but the board continues to malfunction, there is a strong possibility that the trouble is with the associated wiring or switches.

5.6 SIGNAL CHANNEL

All of the gain switching in the Model 124A is done by means of relays. In the following checks, the various amplifiers are isolated by appropriately selecting the Sensitivity and Function. By applying a suitable signal and checking at critical points, a malfunctioning amplifier can be quickly identified. Table V-1, which lists the gain and energized relays for all possible combinations of Sensitivity and PSD Function, is provided as a convenient reference.

Note that overload level signals are applied at various points throughout the following procedure to assure that the signal level at the output of the early amplifiers will be above the noise floor. Anytime the applied signal is greater than the selected sensitivity, overload is a possibility, and the operator should not be concerned if the Overload light glows during such a measurement. However, if the Overload light should glow with normal signal levels applied, a malfunction is indicated, and it should be corrected before proceeding.

5.6A PREAMPLIFIER

NOTE: If the preamplifier to be used is a Model 118 or Model 185, set the Calibrator Output level to 100 μV and the Sensitivity to 1 mV. In the case of a type 117 Preamplifier, or a Model 116 or 119 operated direct, both the Sensitivity and the Calibrator Output level should be set to 1 mV. For all preamplifiers, the Function switch should be set to NORMAL.

(1) Connect a cable from the Calibrate Output to the "A" Preamplifier Input.

(2) Monitor the signal at TP1000 (green testpoint) on the Signal board. The observed signal should be an 11 mV pk-pk square wave at 399 Hz, indicating that the preamplifier gain is five (fifty for a Model 118). **NOTE**: A square wave with an rms value (fundamental frequency component only) of 1 mV has a pk-pk amplitude of 2.22 mV. Therefore, the total signal amplitude at the output of the Preamplifier is 2.22 x 5 = 11 mV. This "X2.22" factor must be taken into account throughout the entire procedure. The operator is again cautioned not to spend an undue amount of effort convincing himself that the observed signals comply with the text descriptions down to the last decimal place. In most instances of malfunction, the signal discrepancy will be so large as to leave no doubt.

If the signal is as indicated, the operator can conclude that the Preamplifier is functioning properly in its gain-of-five mode (gain-of-fifty for Model 118 or Model 185). If the signal is incorrect or missing, the trouble probably is in the Preamplifier and the operator can proceed to the schematic for his particular Preamplifier if he wishes to troubleshoot further. As can be seen from the Signal board schematic on page VI-9, the signal is actually being monitored after relay K1008 on the Signal board. Hence, if the signal is missing, it might be worth checking for the signal on coupling capacitor C1004 to isolate the relay. Also, take a moment to be sure that the Calibrator Output signal is normal, that is, a square wave with a pk-pk amplitude of 2.22 times the selected Calibrator Output level.

(3) Change the Sensitivity setting to 500 μV, thereby energizing Preamplifier relay K100 and increasing the

AC RELAYS AND GAIN

SELECTED SENSITIVITY			PREAMPLIFIER		SIGNAL AMP		INTERMEDIATE AMP		MIXER BD AC		TOTAL
FUNCTION SWITCH SETTING			ENERGIZED	PA	ENERGIZED	SA	ENERGIZED RELAY	IA	ENERGIZED	AC	AC GAIN
LOW DRIFT	NORMAL	HI DYN RNG	RELAY	GAIN[1]	RELAY	GAIN		GAIN	RELAY	GAIN[2]	
100 nV			K100	X50	K1008	X10	K2007, K2003, K2005	X100	K3101	X11	5.5×10^5
200 nV			"	"	"	"	K2008, " "	X50	"	"	2.7×10^5
500 nV			"	"	"	"	K2009, " "	X20	"	"	1.1×10^5
1 μV	100 nV		"	"	"	"	K2007, " "	X100	"	"	5.5×10^5
2 μV	200 nV		"	"	"	"	K2008, " "	X50	"	"	2.7×10^5
5 μV	500 nV		"	"	"	"	K2009, " "	X20	"	"	1.1×10^5
10 μV	1 μV	100 nV	"	"	"	"	K2007, K2004, "	X10	"	"	5.5×10^4
20 μV	2 μV	200 nV	"	"	"	"	K2008, " "	X5	"	"	2.7×10^4
50 μV	5 μV	500 nV	"	"	"	"	K2009, " "	X2	"	"	1.1×10^4
100 μV	10 μV	1 μV	"	"	"	"	K2007, " K2006	X1	"	"	5.5×10^3
200 μV	20 μV	2 μV	"	"	"	"	K2008, " "	X.5	"	"	2.7×10^3
500 μV	50 μV	5 μV	"	"	"	"	K2009, " "	X.2	"	"	1.1×10^3
1 mV	100 μV	10 μV	"	"	"	"	K2007, " "	X1	K3102	X1.1	5.5×10^2
2 mV	200 μV	20 μV	"	"	"	"	K2008, " "	X.5	"	"	2.7×10^2
5 mV	500 μV	50 μV	"	"	"	"	K2009, " "	X.2	"	"	1.1×10^2
10 mV	1 mV	100 μV	NONE	X5	"	"	K2007, " "	X1	"	"	55
20 mV	2 mV	200 μV	"	"	"	"	K2008, " "	X.5	"	"	27
50 mV	5 mV	500 μV	"	"	"	"	K2009, " "	X.2	"	"	11
100 mV	10 mV	1 mV	"	"	K1009	X1	K2007, " "	X1	"	"	5.5
200 mV	20 mV	2 mV	"	"	"	"	K2008, " "	X.5	"	"	2.7
500 mV	50 mV	5 mV	"	"	"	"	K2009, " "	X.2	"	"	1.1
	100 mV	10 mV	"	"	"	"	K2007, " "	X1	"	"	5.5
	200 mV	20 mV	"	"	"	"	K2008, " "	X.5	"	"	2.7
	500 mV	50 mV	"	"	"	"	K2009, " "	X.2	"	"	1.1
		100 mV	"	"	"	"	K2007, " "	X1	"	"	5.5
		200 mV	"	"	"	"	K2008, " "	X.5	"	"	2.7
		500 mV	"	"	"	"	K2009, " "	X.2	"	"	1.1

DC GAIN

MIXER BOARD DC						
LOW DRIFT		NORMAL		HI DYN RNG		
RELAYS	GAIN	RELAYS	GAIN	RELAYS	GAIN	
K3202	200	READ CORRESPONDING SELECTED SENSITIVITY FROM COLUMN AT LEFT SIDE OF TABLE				
"	"					
"	"					
NONE	20	K3202	200			
"	"	"	"			
"	"	"	"			
"	"	"	"	K3201, K3202	2000	
"	"	"	"	"	"	
"	"	"	"	"	"	
"	"	"	"	"	"	
"	"	"	"	"	"	
"	"	"	"	"	"	
"	"	"	"	"	"	
"	"	"	"	"	"	
"	"	"	"	"	"	
"	"	"	"	"	"	
"	"	"	"	"	"	
"	"	"	"	"	"	
"	"	"	"	"	"	
"	"	"	"	"	"	
"	"	"	"	"	"	
"	"	"	"	"	"	
"	"	"	"	"	"	
"	"	NONE	20	K3202	200	
"	"	"	"	"	"	
"	"	"	"	"	"	
"	"	"	"	NONE	20	
"	"	"	"	"	"	
"	"	"	"	"	"	

TOTAL GAIN

GAIN	SENS
TOTAL GAIN DOES NOT VARY WITH SELECTED FUNCTION (LD, N, HDR).	MATCH TOTAL GAIN TO SELECTED SENSITIVITY FROM THIS COLUMN
1.1×10^8	100 nV
5.5×10^7	200 nV
2.2×10^7	500 nV
1.1×10^7	1 μV
5.5×10^6	2 μV
2.2×10^6	5 μV
1.1×10^6	10 μV
5.5×10^5	20 μV
2.2×10^5	50 μV
1.1×10^5	100 μV
5.5×10^4	200 μV
2.2×10^4	500 μV
1.1×10^4	1 mV
5.5×10^3	2 mV
2.2×10^3	5 mV
1.1×10^3	10 mV
5.5×10^2	20 mV
2.2×10^2	50 mV
1.1×10^2	100 mV
5.5×10^1	200 mV
2.2×10^1	500 mV

[1] STATED PREAMPLIFIER GAIN FIGURES APPLY FOR MODEL 117 PREAMP. AND FOR MODEL 116 OR 119 OPERATED DIRECT. WITH MODEL 118 OR MODEL 185 PREAMP. GAIN WILL BE FACTOR OF 10 HIGHER. IF TRANSFORMER COUPLING IS USED, TRANSFORMER TURNS RATIO MUST BE TAKEN INTO ACCOUNT AS WELL

[2] GAINS ARE TEN PERCENT HIGH TO COMPENSATE FOR THE FACT THAT THE MIXER IS AVERAGE RESPONDING AS OPPOSED TO RMS RESPONDING

Table V-1. GAIN AND RELAY SWITCHING FOR THE MODEL 124A

preamplifier gain to X50 (X500 for a Model 118 or a Model 185). The amplitude of the signal at TP1000 should increase to 111 mV pk-pk, reflecting the increased gain. If this check is normal, one can reasonably assume that the preamplifier is functioning normally, at least with respect to providing the proper gain. Noise and common-mode rejection checks are beyond the scope of this procedure. If this check does not give a normal indication, but step 2 does give a normal indication, relay K100 should be suspected.

5.6B SIGNAL AMPLIFIER

The Signal Amplifier gain, measured from the output of the Preamplifier to the output of the Signal Amplifier, is either X10 (K1008 energized) or X1 (K1009 energized). Note that the gain of the amplifier circuit is always X10. The factor of ten gain reduction achieved when K1009 is energized and K1008 is de-energized is accomplished by reducing the amplitude of the signal applied to the amplifier input with a relay controlled attenuator.

(1) Transfer the oscilloscope to TP1001 (violet testpoint). The observed signal should be a 1.11 V pk-pk square wave, indicating that the Signal Amplifier has a gain of ten. If the signal is normal, chances are the amplifier circuit is functioning normally. If the signal is absent, the problem could lie with the amplifier circuit or with K1008, the energized input relay.

(2) Set the Sensitivity switch to "10 mV". The signal amplitude should decrease to 11 mV pk-pk, indicating that the gain of both the Preamplifier and Signal Amplifier went down by a factor of ten. If the signal is as indicated, chances are that the Signal Amplifier Input Attenuator is functioning normally. If an incorrect reading is obtained, K1009, the input attenuator relay energized with this Sensitivity switch setting, should be checked.

(3) Set the Signal Mode switch to BANDPASS and the Signal Channel Frequency controls to 3.99, X100 (399 Hz). Then set the Signal Q switch to "10". The observed signal at TP1001 should now be sinusoidal with an amplitude of 14 mV pk-pk. It may be necessary to slightly adjust the third Signal Frequency dial to obtain the indicated amplitude. The amplitude drops off sharply if the dial is set high or low. If the indicated effects are observed, the selective amplifier circuits of the Signal Amplifier are functioning normally.

5.6C INTERMEDIATE AMPLIFIER

There are two amplifiers on the Intermediate Amplifier board, each with a nominal gain of ten. The relay switching on the board, controlled by the Sensitivity switch, actuates various attenuators so that the overall board gain varies from X100 to X0.2, according to the Sensitivity switch position. Not all possible gains are checked. Instead, each decade is checked, and also the X1, X0.5, X0.2 sequence within one decade. This is sufficient to check all of the relays as well as the amplifiers. Note from the schematic on page VI-13 that the front-panel Sensitivity Adjustment affects the gain of the second amplifier and hence the overall gain of the Intermediate Amplifier. Hence, it may be necessary to change the setting of this adjustment to obtain signals of the indicated level. However, once set, the control setting should not have to be changed again, at least not for the remainder of the Intermediate Amplifier checks.

(1) Set the Sensitivity switch to 100 nV and the Function switch to NORMAL. Then set the Calibrator output to "2 μV" (200 nV with a Model 118 or 185).

(2) Monitor the signal at TP2000 (green testpoint). The observed signal should be a 399 Hz sinewave with a pk-pk amplitude of 0.28 V. It is perfectly normal for this signal to be obscured by noise. If this signal is as indicated, one can assume that both amplifiers on the Intermediate Amplifier board are functioning normally. If the signal is not normal, the problem could be with one of the two amplifier circuits or with one of the relays. With this combination of Sensitivity and Function, the energized relays are K2003, K2005, and K2007.

(3) Set the Sensitivity to "1 μV" and the Calibrate Output to 20 μV (2 μV with a Model 118 or 185). The observed signal should still have the same amplitude (the gain reduction is exactly compensated by the increased calibrator output), but the noise should have gone down by a factor of ten. The only relay change between this step and the preceding one is that K2004 is now energized and K2003 de-energized.

(4) Set the Sensitivity switch to 10 μV and the Calibrate Output to 200 μV (20 μV with a Model 118 or 185). As in the preceding step, the decrease in gain is compensated by the increase in calibrator output and the amplitude of the observed signal should remain constant (0.28 V pk-pk). A further factor of ten reduction in the observed noise will take place. The relay state changes are that K2006 is now energized and K2005 is de-energized.

(5) Set the Sensitivity switch to 2 mV and the Calibrate Output to 2. mV. The amplitude of the observed signal should decrease to 0.14 V pk-pk, indicating the gain reduction which occurs as K2007 is de-energized and K2008 is energized.

(6) Set the Sensitivity switch to 5 mV. The amplitude of the observed signal should decrease to 57 mV pk-pk, reflecting the gain reduction which occurs as K2008 drops out and K2009 is energized.

All of the Intermediate Amplifier board relays have now been checked. If proper operation to this point was obtained, the Intermediate Amplifier board can be presumed to be functioning normally.

5.6D MIXER BOARD AC GAIN

The Mixer board ac gain is either X11 or X1.1 as determined by attenuator controlling relays K3101 and

K3102. The gains are made 10% "high" to compensate for the fact that, although the instrument reads out in rms, the Mixer is average responding.

(1) Set the Sensitivity switch to 50 μV and the Calibrate Output to 500 μV (50 μV with a Model 118 or 185). Then monitor the signal at the front-panel SIGNAL MONITOR connector. The observed signal should have a pk-pk amplitude of 28.3 mV. This signal will be noisy.

(2) Set the Sensitivity switch to 100 μV and increase the Calibrator Output to 1 mV (100 μV with a Model 118 or 185). The observed signal should remain the same in amplitude (gain decrease compensated by increased calibrator output) but the noise should decrease.

If the indicated signal levels were observed, one can assume that the ac amplifier portion of the Mixer circuitry is functioning normally.

5.6E MIXER SCHMITT TRIGGERS

Connect the oscilloscope to TP3002. One should observe a 400 Hz square wave with the lower level at −12 V and the upper one at 0 V. If this waveform is as indicated, the Mixer Schmitt Triggers are probably working normally. If the waveform is absent, check for a 2.8 V pk-pk sinewave at C3005 before concluding there is a problem in the Schmitt Trigger circuit. Note from the schematic on page VI-19 that the blue testpoint gives access to only one "half" of the complementary Schmitt Trigger output. The complement (collector of Q3003) could be checked at R3112 or at R3110.

5.6F MIXER CIRCUIT

(1) Set both the Sensitivity and Calibrate switches to 5 mV (Calibrate Output to 500 μV with a Model 118), and the Function switch to LO DRIFT.

(2) Monitor the signal at the junction of R3202 and R3204. Then set the Phase switch to 270°. The observed signal should be a full-wave rectified sinewave, although a slight adjustment of the Phase dial may be required to obtain this waveform. The amplitude of the half-waves should be about 80 mV relative to a 0 V baseline. If the signal is as indicated, the Mixer circuit is probably functioning normally.

5.6G DC AMPLIFIERS

With Sensitivity switch settings from 1 μV through 5 mV, the dc gain is determined solely by the position of the Function switch. In LO DRIFT it is X20, in NORMAL it is X200, and in HI DYNAMIC RANGE it is X2000. For a given sensitivity, changing the function does not change the output dc level because the ac gain varies as well as to keep the overall instrument gain constant. Thus the dc amplifiers are checked simply by monitoring the dc output and observing that it does not change as the Function switch is rotated through its three PSD positions.

(1) Adjust the Phase dial and third Signal Frequency dial for maximum panel meter deflection. Then set the screwdriver adjustable Sensitivity Adj. control for exactly full-scale panel meter deflection. If this cannot be done, there is probably a malfunction in one of the dc amplifiers or in the meter. The meter can be eliminated by checking the dc level at the front-panel Function Out connector. Ten volts corresponds to full-scale panel meter deflection.

(2) At this point, the Function switch should still be set to LO DRIFT. Successively set it to NORMAL and then to HI DYNAMIC RANGE. The panel meter should continue to indicate full scale ±3%. If it does, one can reasonably assume that the dc amplifiers are functioning normally. If necessary, one could check the output of the first dc amplifier separately. This is most easily done at R3228, the emitter resistor of Q3203 (schematic on page VI-22). The voltage there should be −1.5 V ±0.5 V. The output of the second amplifier should, of course, be +10 V.

(3) Check the ACVM function by setting the Function switch to ACVM. The meter indication should remain unchanged. Note that ac voltmeter operation is achieved simply by taking the drive to the Mixer Schmitt Trigger from the Signal Channel instead of from the Reference Channel. No new circuits are activated.

If the instrument has passed all tests to this point, but the unit is still malfunctioning in some way not revealed by these tests, then the problem is beyond the scope of this troubleshooting procedure and the operator should contact the factory or one of its authorized representatives for advice on how to proceed.

5.7 NOISE CHECKS

These checks allow the operator to determine whether the internally generated noise in his instrument is normal. Note that these checks vary according to the type of preamplifier used.

(1) Set the front-panel controls as follows. **NOTE**: If preamplifier is a Model 184, go directly to step 10.

Input Selector (Models 116, 117, 118, & 119): A
Transformer/Direct switch (Models 116 & 119): DIRECT
Ground Isolation (Model 185): IN
Sensitivity
  Models 116, 117, & 119: 1 mV
  Models 118 & 185: 10 mV
Signal Mode: BANDPASS
Signal Frequency dials: 4.05
Signal Range: X100
Signal "Q": 100
Reference Frequency dials: 4.05
Reference Frequency Range: X100
Phase switch: 270°

Phase dial: 90°
Zero Offset toggle switch: OFF (center position)
Function switch: ACVM
Calibrate switch: 1 mV

(2) Connect a cable from the front-panel CALIBRATE connector to the preamplifier input.

(3) Adjust the right-most Signal Frequency control for peak panel-meter indication.

Time Constant switch to 100 ms (Preamplifier is Model 116, 117, or 119) or to 300 ms (Preamplifier is Model 118 or 185).

(9) Note the pk-pk meter fluctuates about zero over a ten second period. If the preamplifier is a Model 116, 117, or 119, the fluctuations should not exceed ±25% of meter full scale. If the preamplifier is a Model 118 or a Model 185, the fluctuations should not exceed ±80% of meter full scale.

SECTION VI
SCHEMATICS

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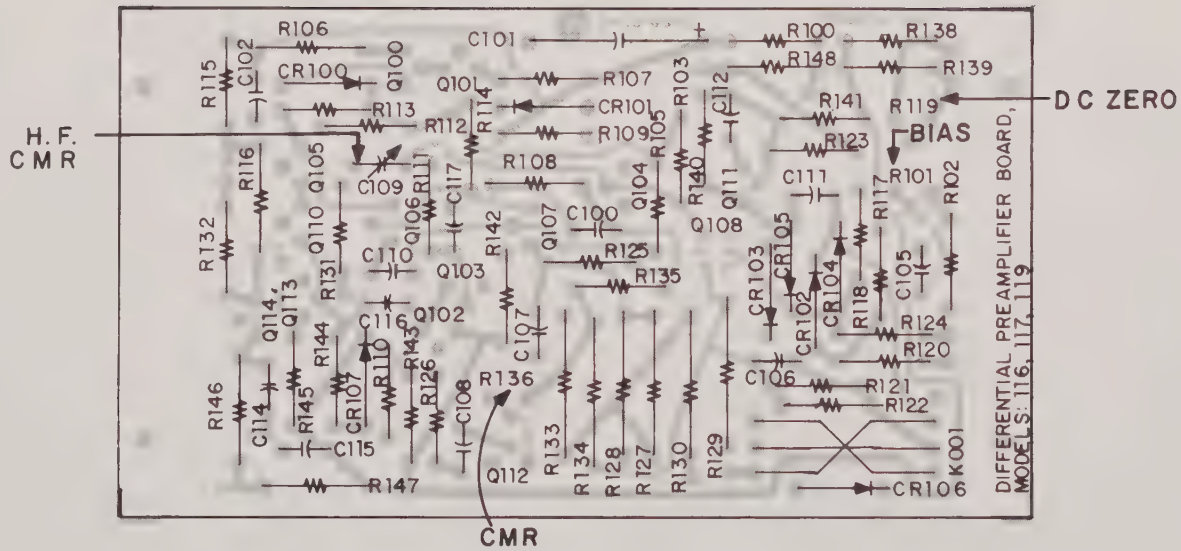
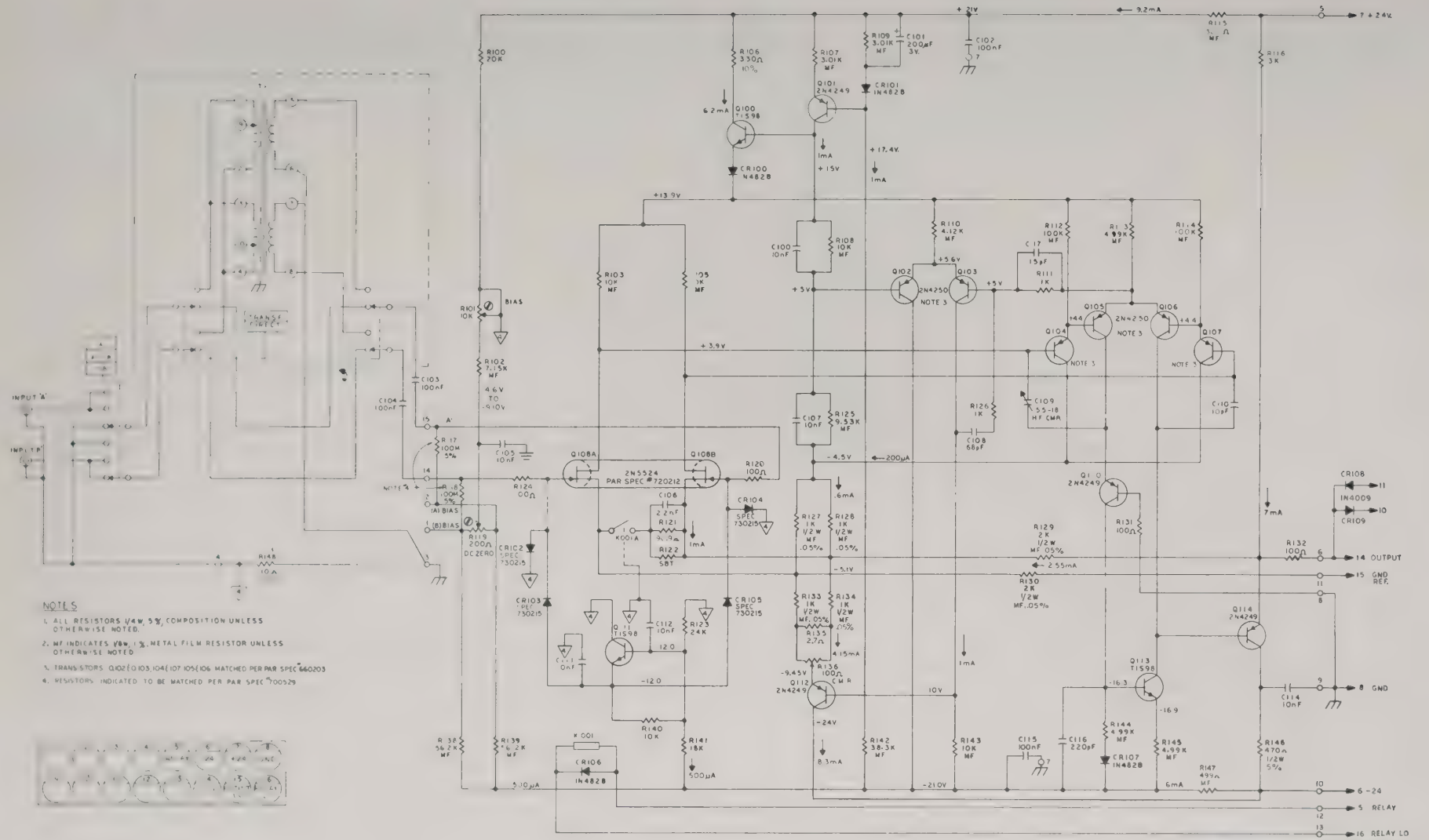


SYMBOLIZATION MASTER

DIFFERENTIAL PREAMPLIFIER BOARD

MODELS: 116, 117, 119



DIFFERENTIAL PREAMPLIFIER
SUB ASSEMBLY 1160-2000-00s

MODEL 116

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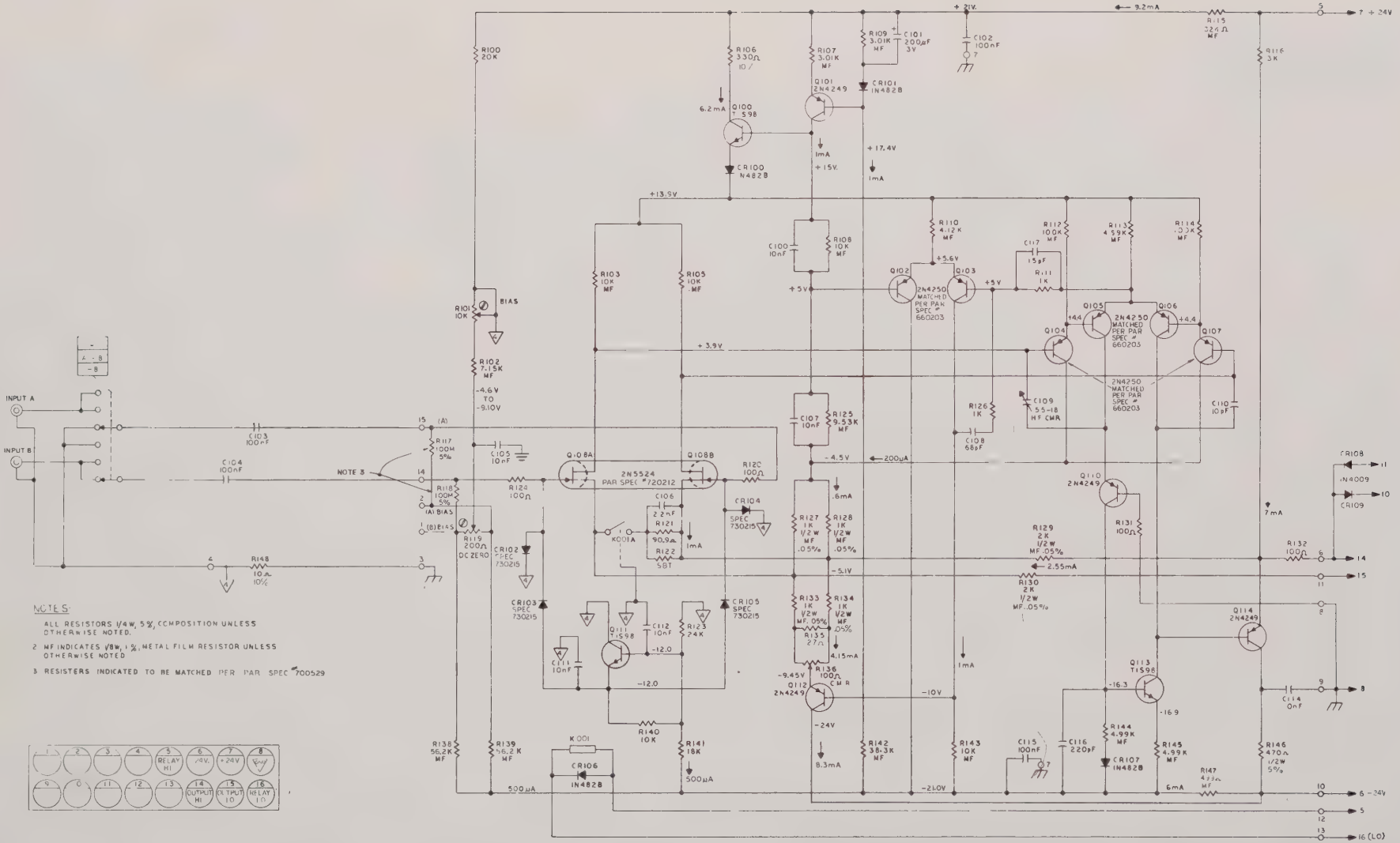
4852 | D | SD | F



DIFFERENTIAL PREAMPLIFIER
SUB ASSEMBLY 1160 2000 00s

MODEL 117

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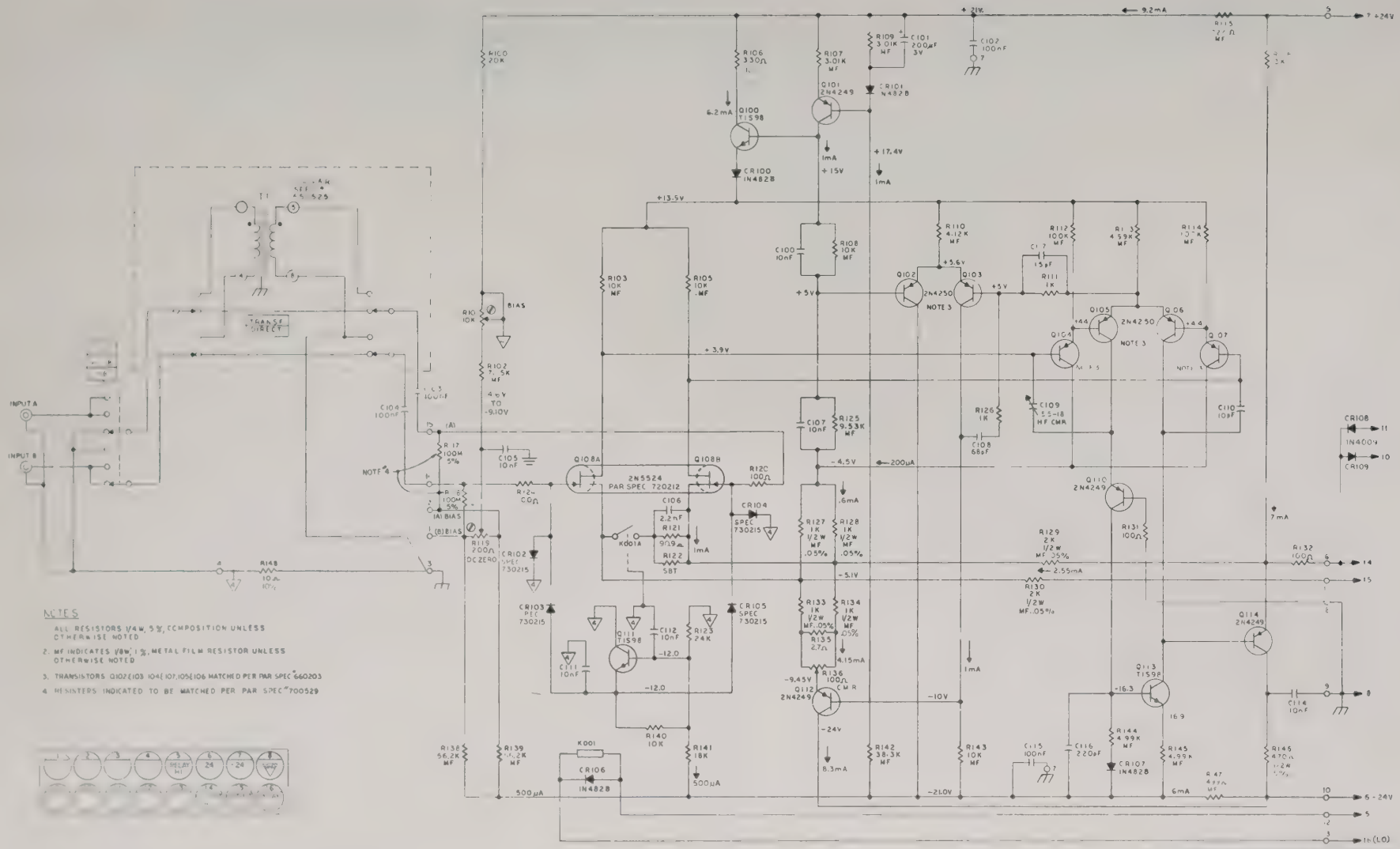
5075 D SD E



DIFFERENTIAL PREAMPLIFIER
R ASSEMBLY 1160 2000 00S

MODEL 119

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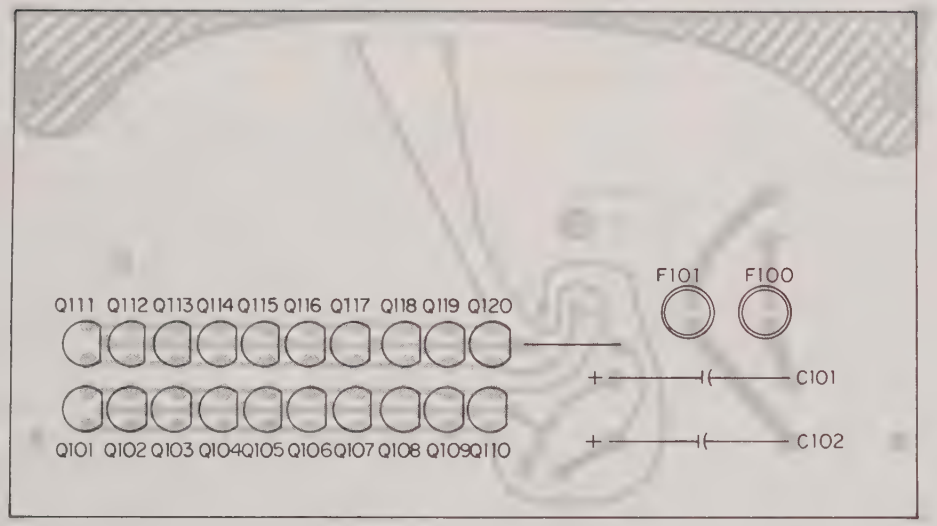
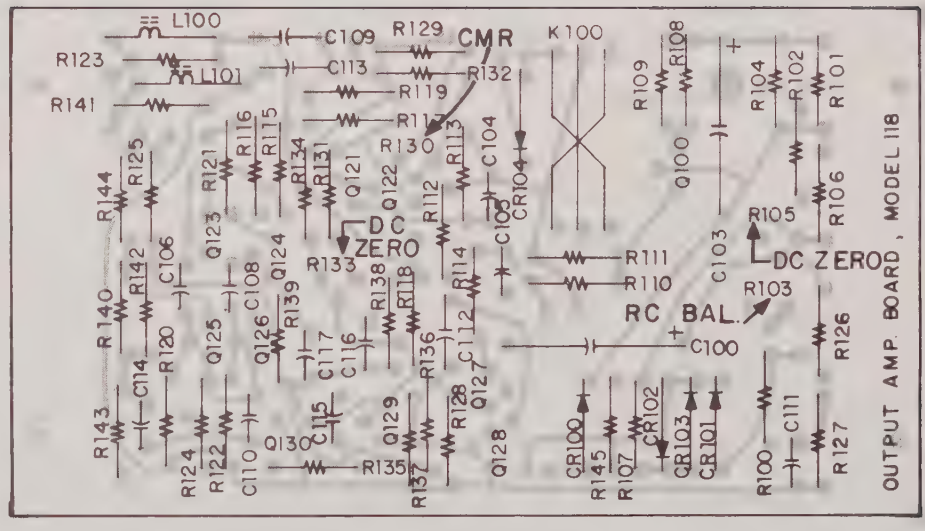
5076 D SD F



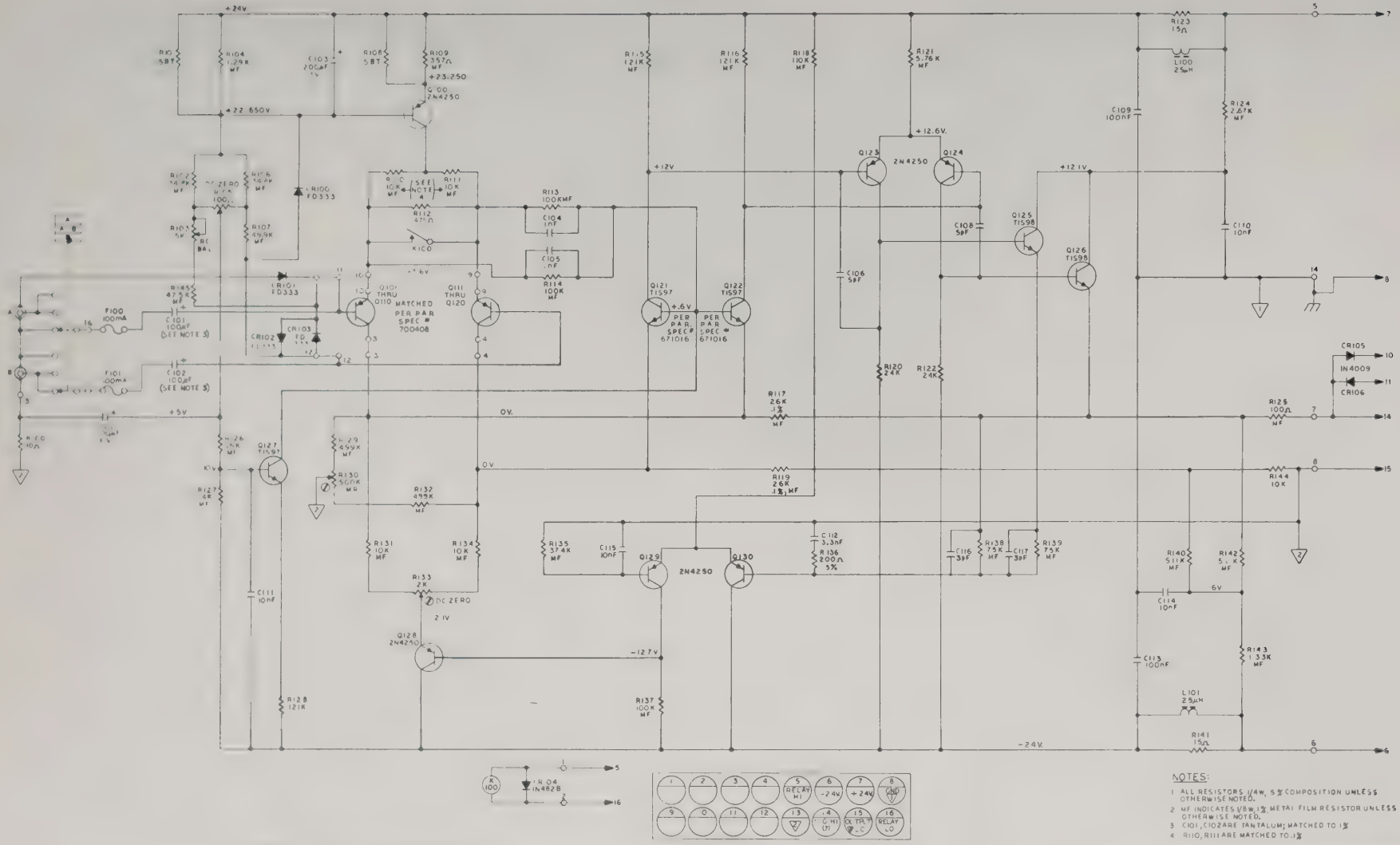
MODEL 118
INPUT AMP BOARD



SYMBOLIZATION MASTER

OUTPUT AMP. BOARD

MODEL 118



MEDIUM Z DIFFERENTIAL PREAMPLIFIER

INPUT AMP. ASSEMBLY 1180-2100-05S
(Q101 THRU Q120; C101 & C102; F100 & F101)
OUTPUT AMP. ASSEMBLY 1180-2000-06S

MODEL 118

NOTES:

1. ALL RESISTORS 1/4W, 5% COMPOSITION UNLESS OTHERWISE NOTED.
2. MF INDICATES 1/8W, 1% METAL FILM RESISTOR UNLESS OTHERWISE NOTED.
3. C101, C102 ARE TANTALUM; MATCHED TO 1%
4. R110, R111 ARE MATCHED TO .1%

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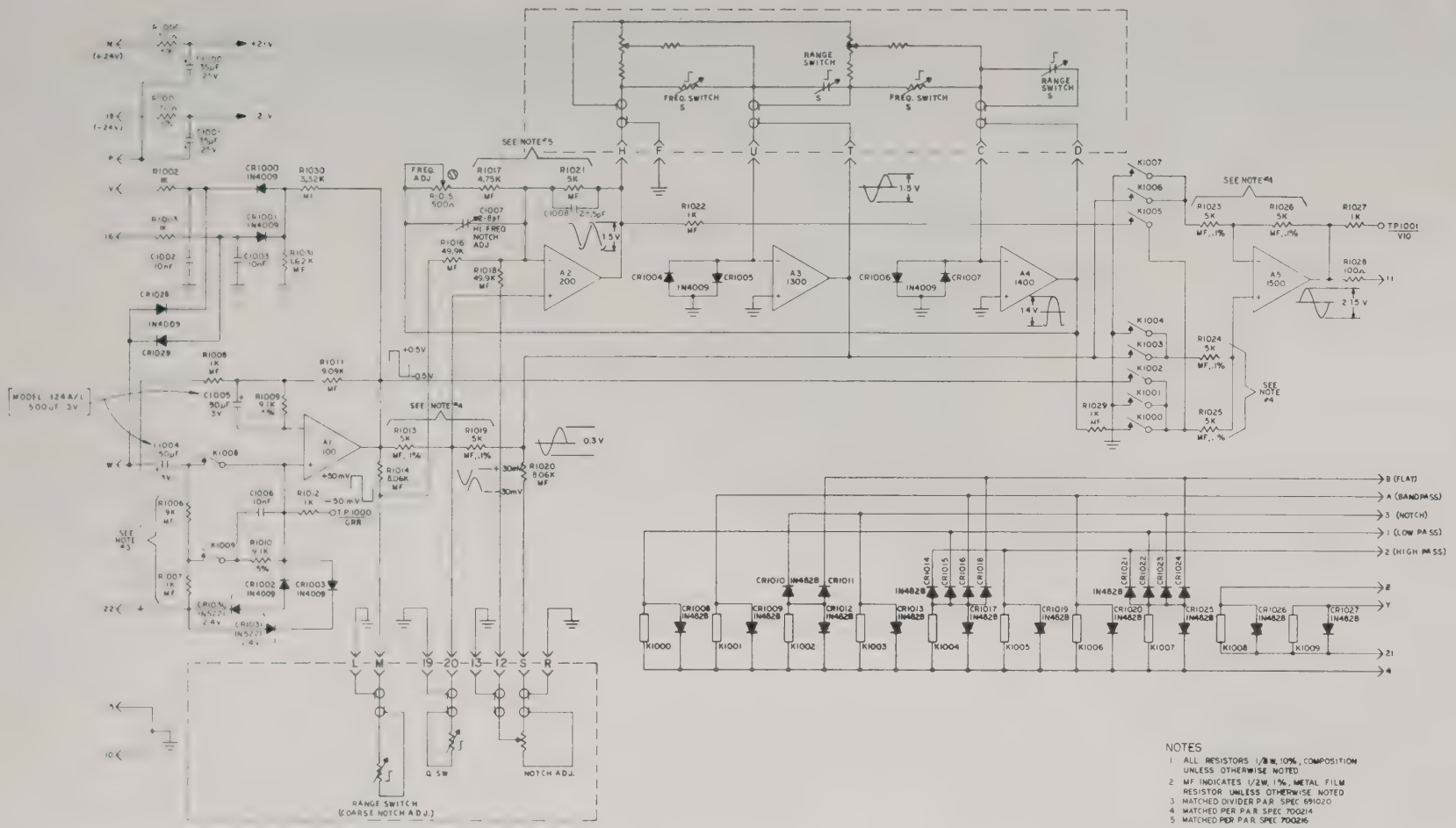
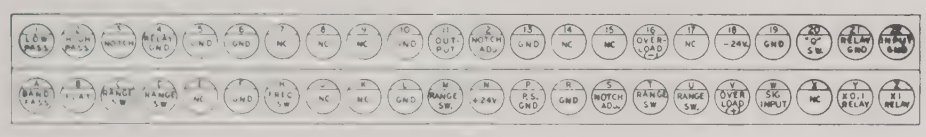
5074 | D | SD | B

SYMBOLIZATION
MODEL 124, 124A
SELECTIVE AMP. BOARD
FAB. # 4971 - MD - C



NOTES

1. ALL RESISTORS 1/2W, 10%, COMPOSITION UNLESS OTHERWISE NOTED
2. MF INDICATES 1/2W, 1%, METAL FILM RESISTOR UNLESS OTHERWISE NOTED
3. MATCHED DIVIDER PAIR SPEC 691020
4. MATCHED PER PAIR SPEC 700214
5. MATCHED PER PAIR SPEC 700216



SIGNAL AMP BD
SUB-ASSEMBLY
*1240-2500-09 (PARTIAL)

MODEL 124 & 124A

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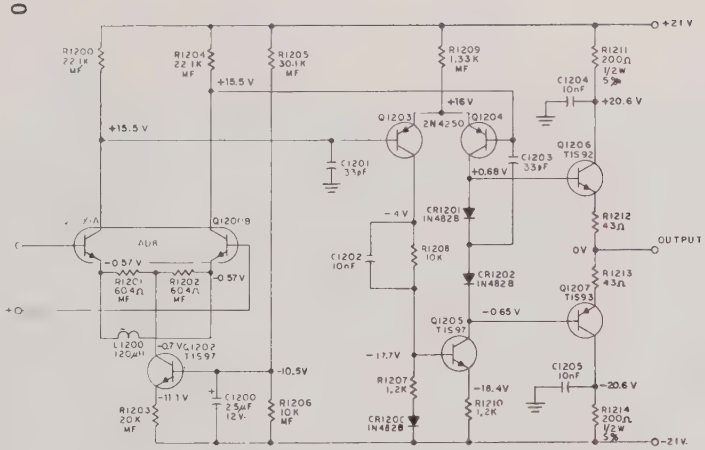
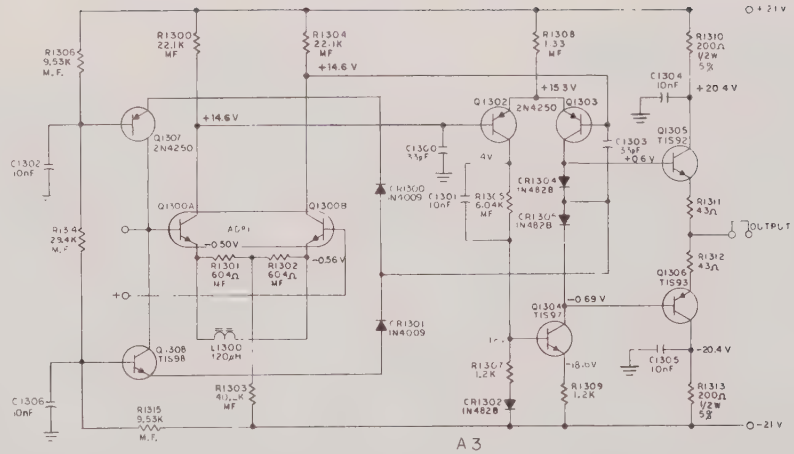
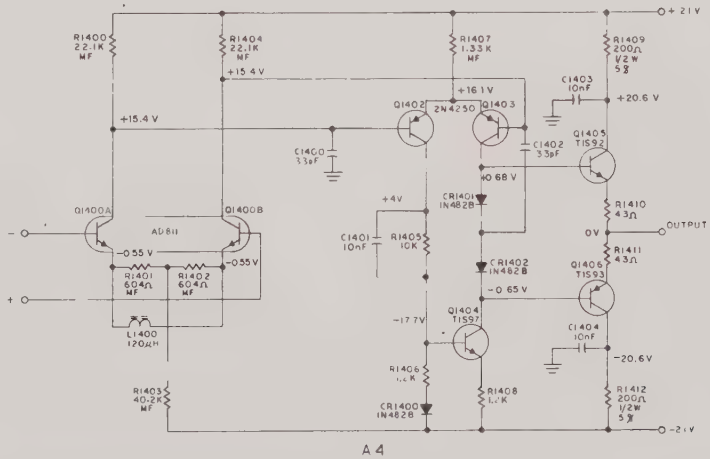


A2



A3



A4

NOTES:
1. ALL RESISTORS ARE 1/4W, 5%, COMPOSITION UNLESS OTHERWISE NOTED.
2. MF INDICATES 1/8 W, 1%, METAL FILM RESISTOR UNLESS OTHERWISE NOTED

OPERATIONAL AMPLIFIERS A2,A3,A4
SIGNAL AMP BD.
SUB-ASSEMBLY #1240-2500-09 (PARTIAL)

MODEL 124 & 124A

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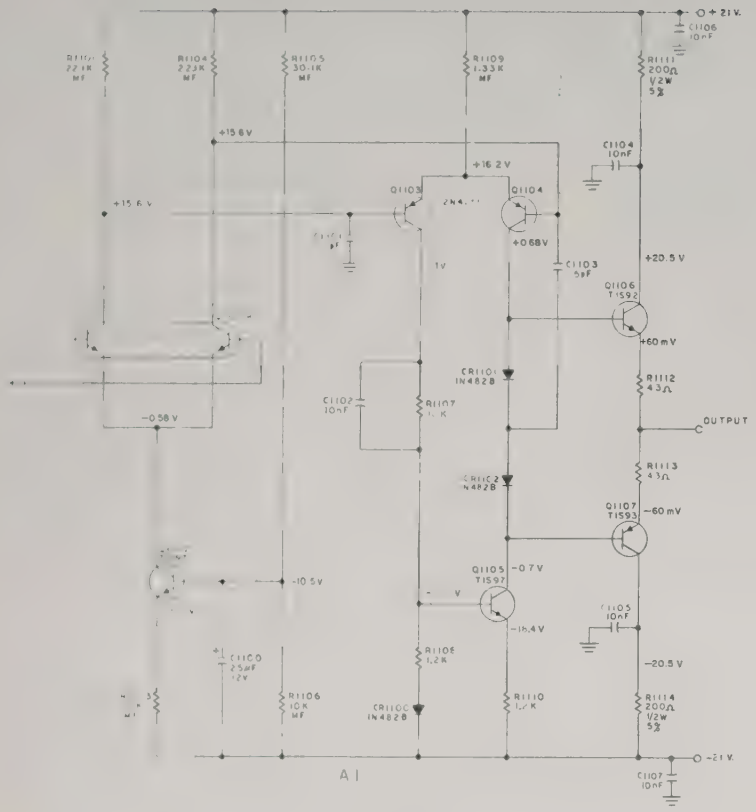

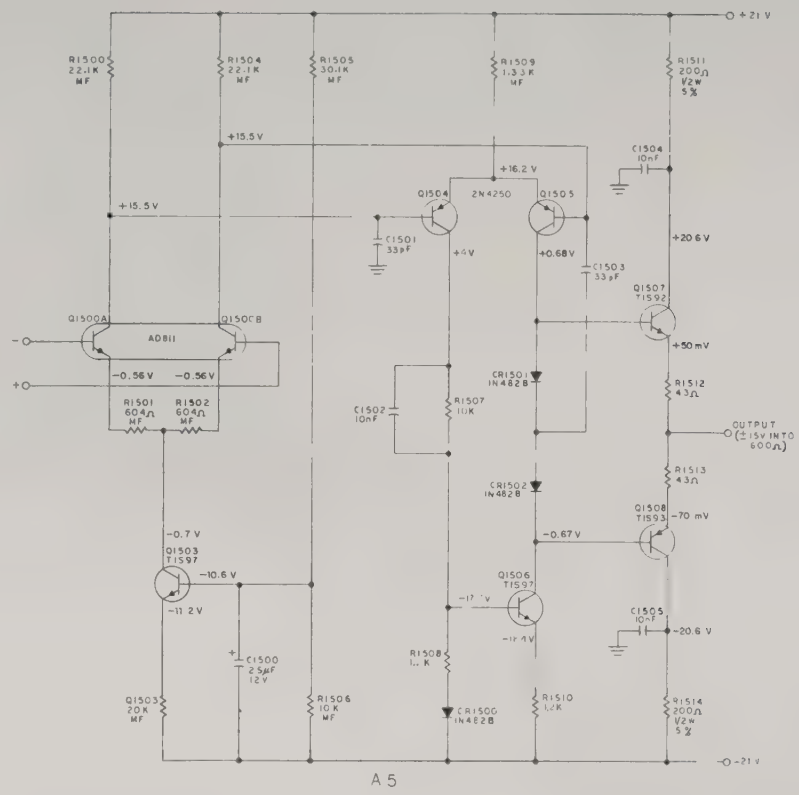

NOTES:
1. ALL RESISTORS ARE 1/4W, 5%, COMPOSITION UNLESS OTHERWISE NOTED.
2. MF INDICATES 1/8W, 1%, METAL FILM RESISTOR UNLESS OTHERWISE NOTED.
3. ALL [illegible] ROUTED TO SHELL OF OUTPUT BNC

OPERATIONAL AMPLIFIERS A1 & A5
SIGNAL AMP BD.
SUB-ASSEMBLY #1240-2500-09 (PARTIAL)

MODEL 124 & 124A

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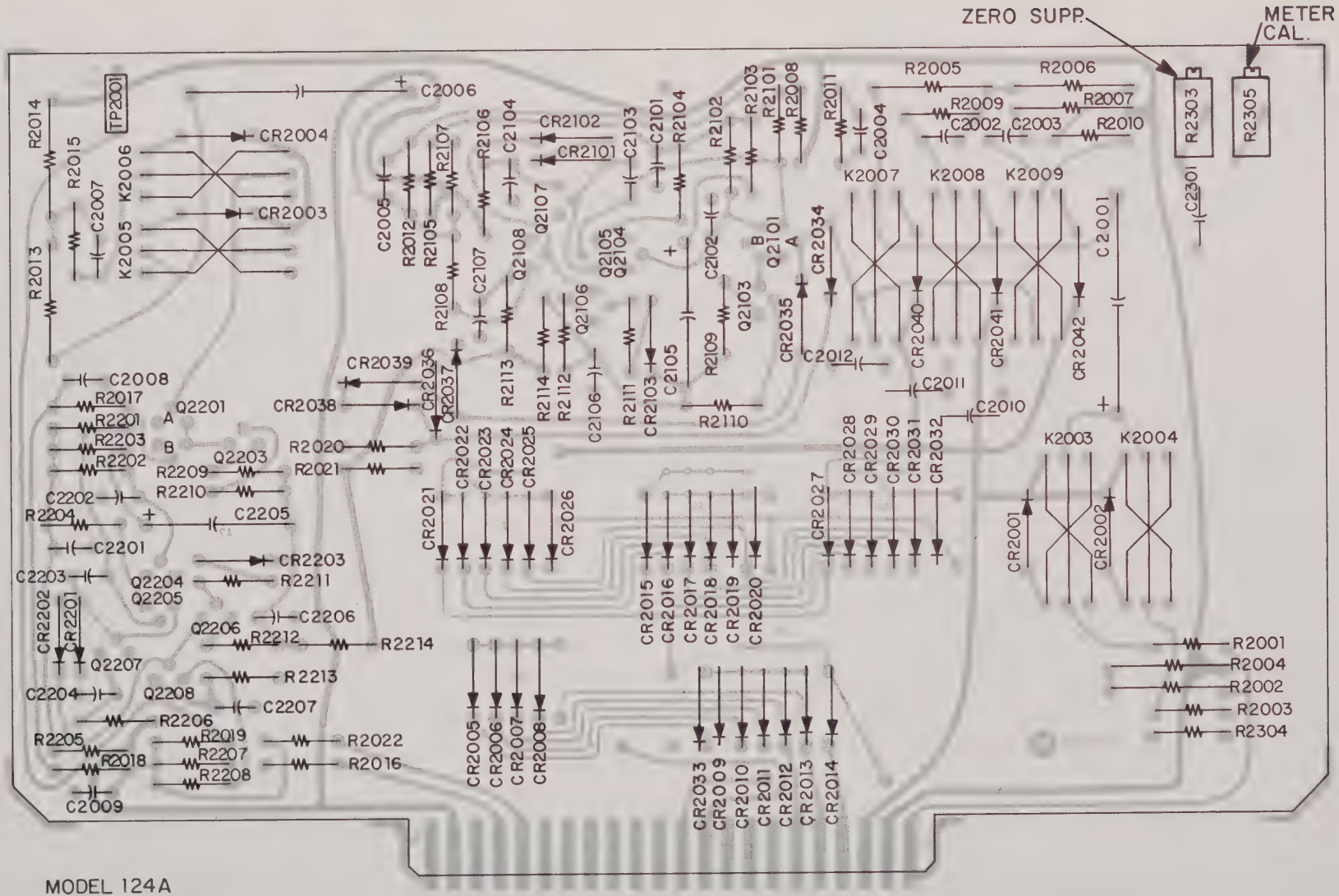
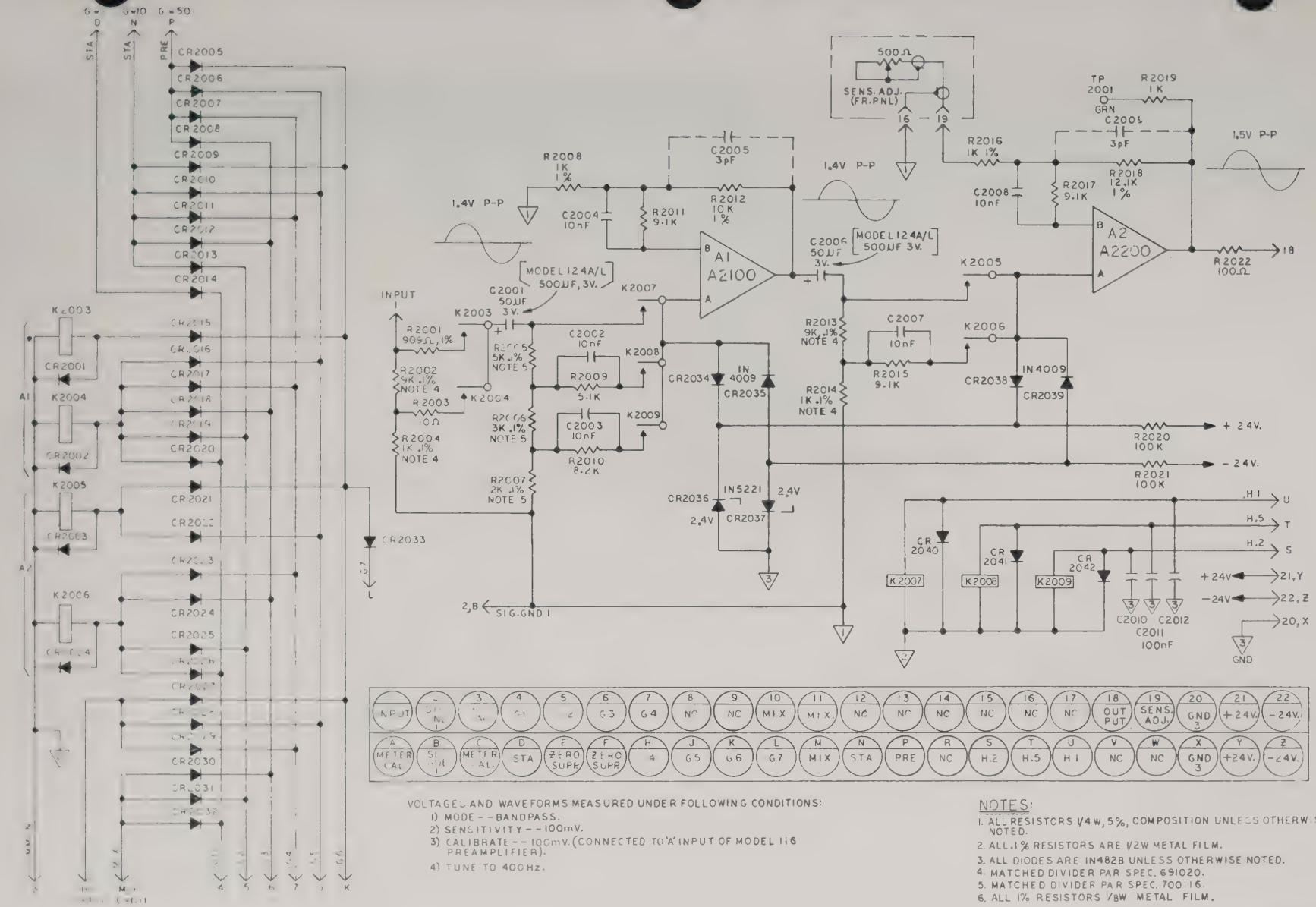
SHEET 3 OF 3 | 4 | D



MODEL 124A
INTERMEDIATE AMP. BOARD



1	2	3	4	5	6	7	8	9	10	11	12	13	14	15	16	17	18	19	20	21	22
INPUT	SIG. GND		G1		G3	G4	NC	NC	MIX	MIX	NC	NC	NC	NC	NC	NC	OUTPUT	SENS. ADJ.	GND 3	+24V.	-24V.
A METER CAL	B SIG. GND 1	C METER CAL	D STA	F ZERO SUPP	F ZERO SUPP	H 4	J G5	K G6	L G7	M MIX	N STA	P PRE	R NC	S H.2	T H.5	U H1	V NC	W NC	X GND 3	Y +24V.	Z -24V.

VOLTAGES AND WAVEFORMS MEASURED UNDER FOLLOWING CONDITIONS:

1) MODE - - BANDPASS.
2) SENSITIVITY - - 100mV.
3) CALIBRATE - - 100mV. (CONNECTED TO 'A' INPUT OF MODEL 116 PREAMPLIFIER).
4) TUNE TO 400Hz.

NOTES:

1. ALL RESISTORS 1/4W, 5%, COMPOSITION UNLESS OTHERWISE NOTED.
2. ALL .1% RESISTORS ARE 1/2W METAL FILM.
3. ALL DIODES ARE IN482B UNLESS OTHERWISE NOTED.
4. MATCHED DIVIDER PAR SPEC. 691020.
5. MATCHED DIVIDER PAR SPEC. 700116.
6. ALL 1% RESISTORS 1/8W METAL FILM.

INTERMEDIATE AMPLIFIER BD

MODEL 124A

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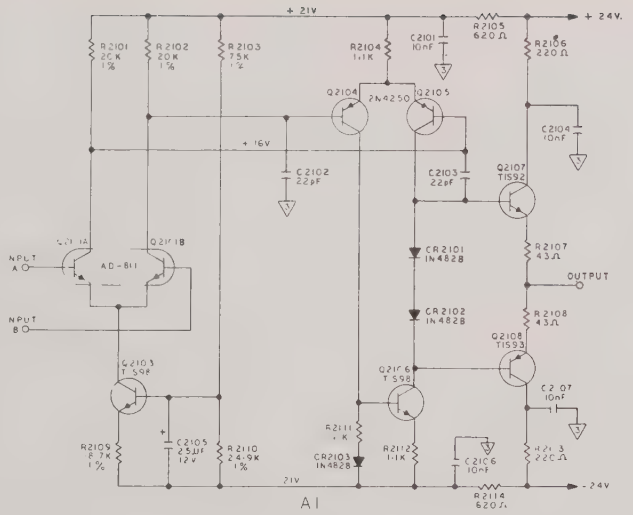

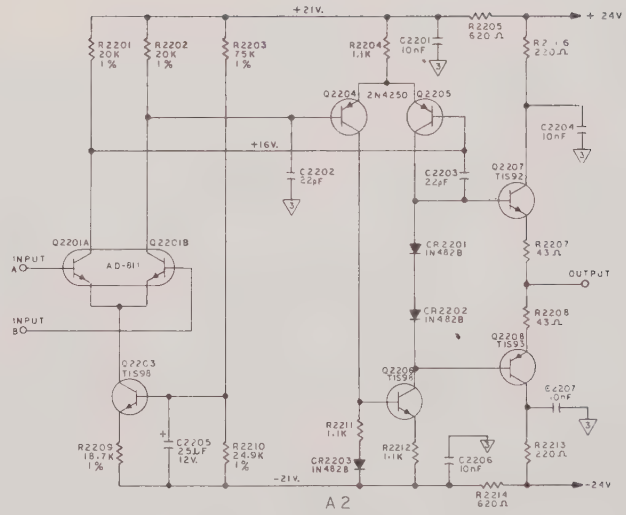

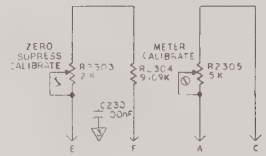

NOTES

SEE SHEET 1

INTERMEDIATE AMPLIFIER BD

MODEL 124A*

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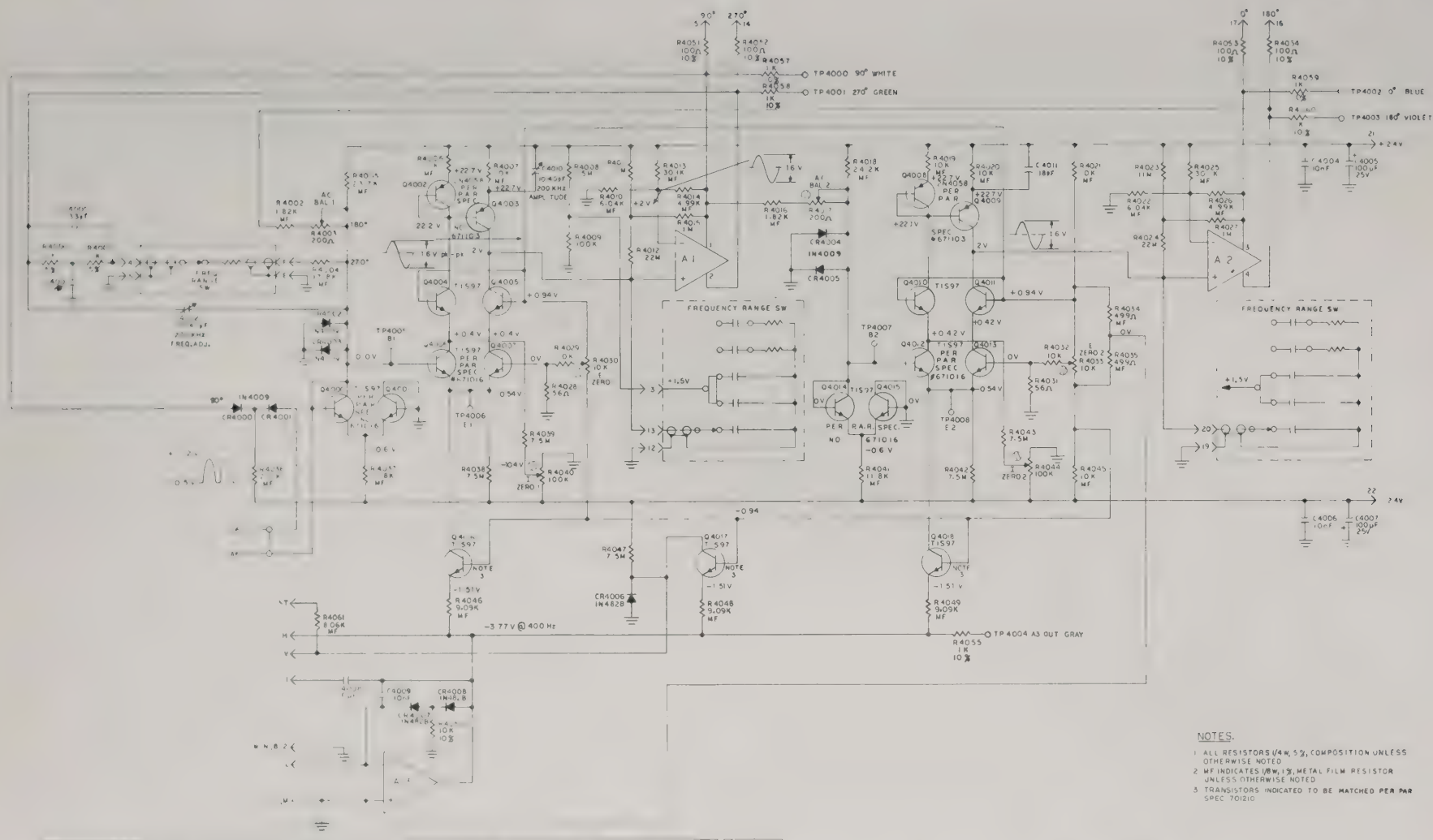

NOTES.

1. ALL RESISTORS 1/4W, 5%, COMPOSITION UNLESS OTHERWISE NOTED
2. MF INDICATES 1/8W, 1% METAL FILM RESISTOR UNLESS OTHERWISE NOTED
3. TRANSISTORS INDICATED TO BE MATCHED PER PAR SPEC 70120

1	2	3	4	5	6	7	8	9	10	11	12	13	14	15	16	17	18	19	20	21	22
MF	GND	+1.5V	FREQ ADJ	GND	NC	NC	NC	NC	NC	GND	GND	TUNING CAPS	0°	90°	180°	270°	N.C.	GND	TUNING CAPS	+24V	-24V

H	J	K	L	M	N	P	R	S	T	U	V	W	X	Y	Z
A3 OUT	MOD	NC	SUM JCT	GND	N.O.	N	N	N	N	NC	REF BACK	GND	N.C.	+24V	-24V

REFERENCE OSCILLATOR BD. MODEL 124[illegible]
SUB-ASSEMBLY #1240-2200-02 (PARTIAL)

SEE 4870-D-SD FOR SHT 2 OF 3
SEE 4870-D-SD FOR SHT 3 OF 3

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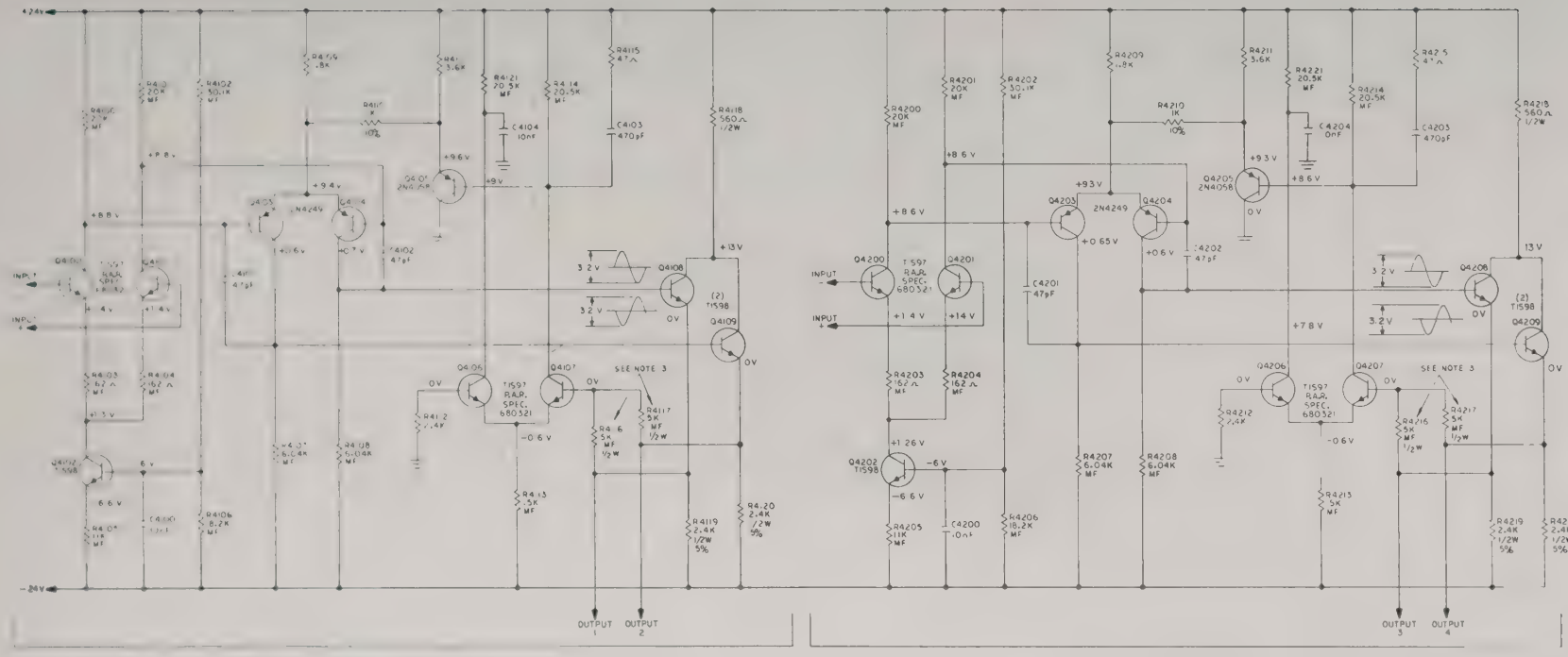
SYMBOLIZATION
MODEL 124, 126
REFERENCE OSCILLATOR BOARD
FAP. # 4980-MD-E



OPERATIONAL AMPLIFIER 'A1'

OPERATIONAL AMPLIFIER 'A2'

NOTES:

1. ALL RESISTORS 1/4W, 5%, COMPOSITION UNLESS OTHERWISE NOTED.
2. MF INDICATES 1/8W, 1%, METAL FILM RESISTOR UNLESS OTHERWISE NOTED.
3. MATCHED PAIR PER SPEC #700214

REFERENCE OSCILLATOR BD. MODEL 124 & 124A
SUB-ASSEMBLY #1240-2200-02 (PARTIAL)

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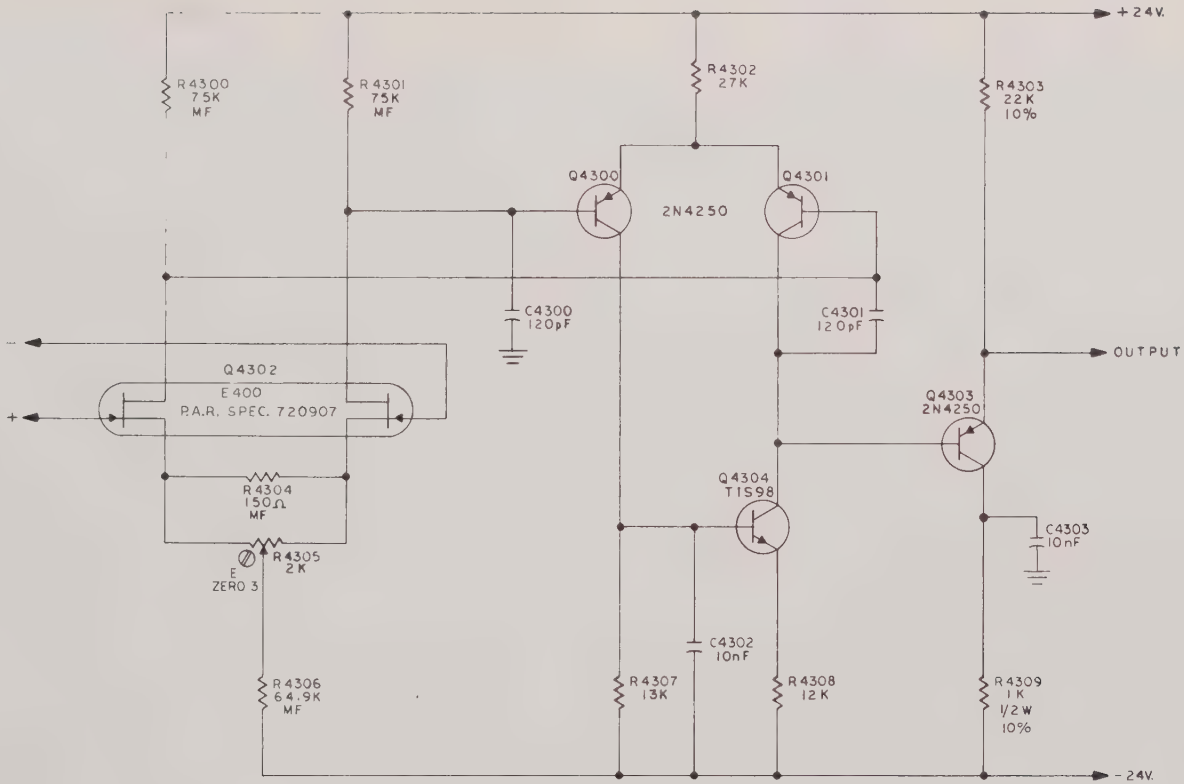

NOTES:

1 ALL RESISTORS 1/4W, 5%, COMPOSITION UNLESS OTHERWISE NOTED.
2 MF INDICATES 1/8W, 1%, METAL FILM RESISTOR UNLESS OTHERWISE NOTED.

OPERATIONAL AMPLIFIER 'A3'

REFERENCE OSCILLATOR BD.
SUB-ASSEMBLY #1240-2200-02 (PARTIAL)

MODEL 124 & 124A

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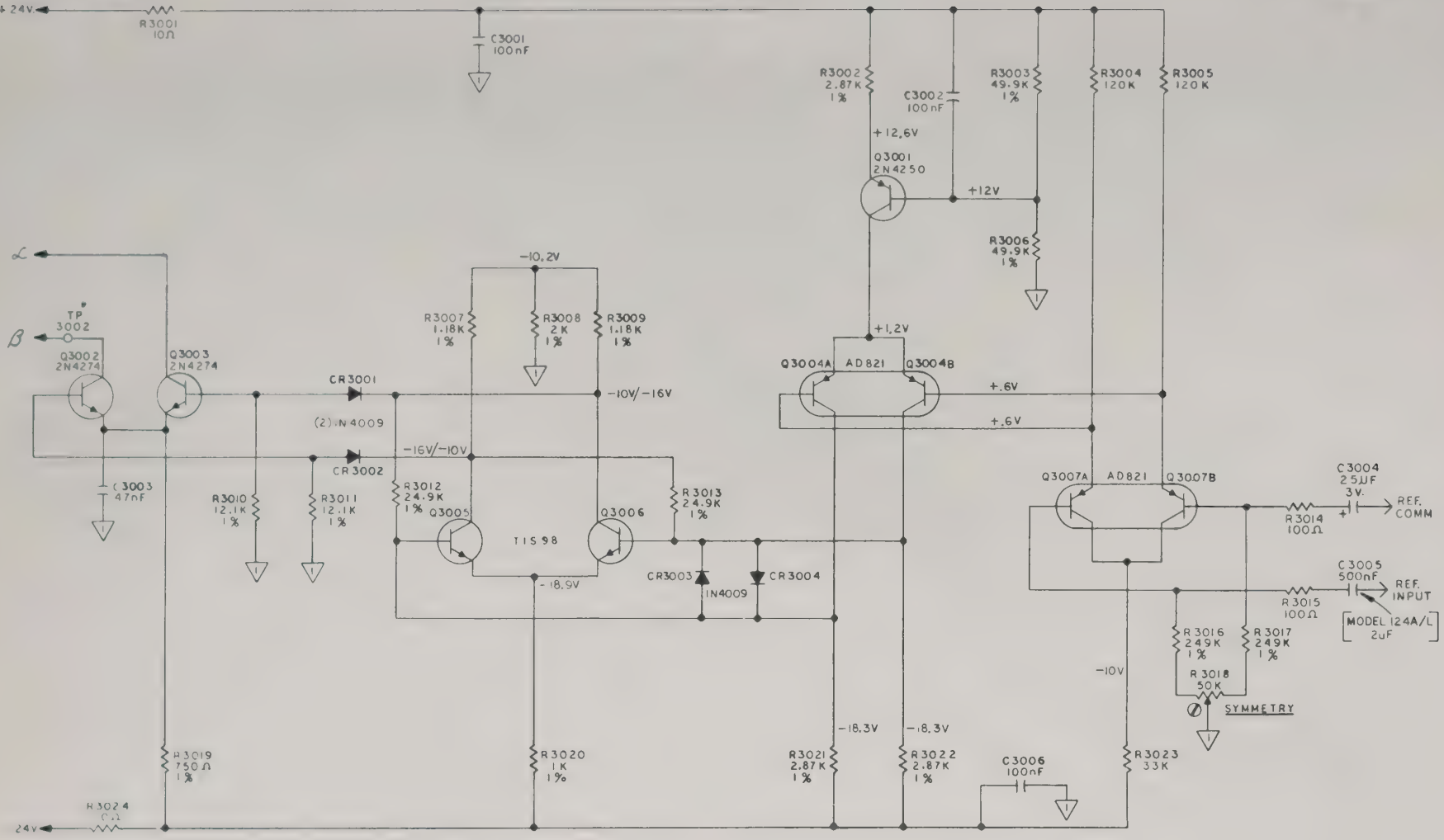
SHEET 3 OF 3 | 4870 | C | SD



NOTES:

1. ALL RESISTORS 1/4W, 5%, COMPOSITION, UNLESS OTHERWISE NOTED.
2. ALL 1% RESISTORS ARE 1/8W METAL FILM UNLESS OTHERWISE NOTED.
3. ALL DIODES ARE IN482B UNLESS OTHERWISE NOTED.

MIXER BOARD

SCHMITT TRIGGER

SUB ASSEMBLY 1241-53-0003S (PARTIAL)

MODEL 124A

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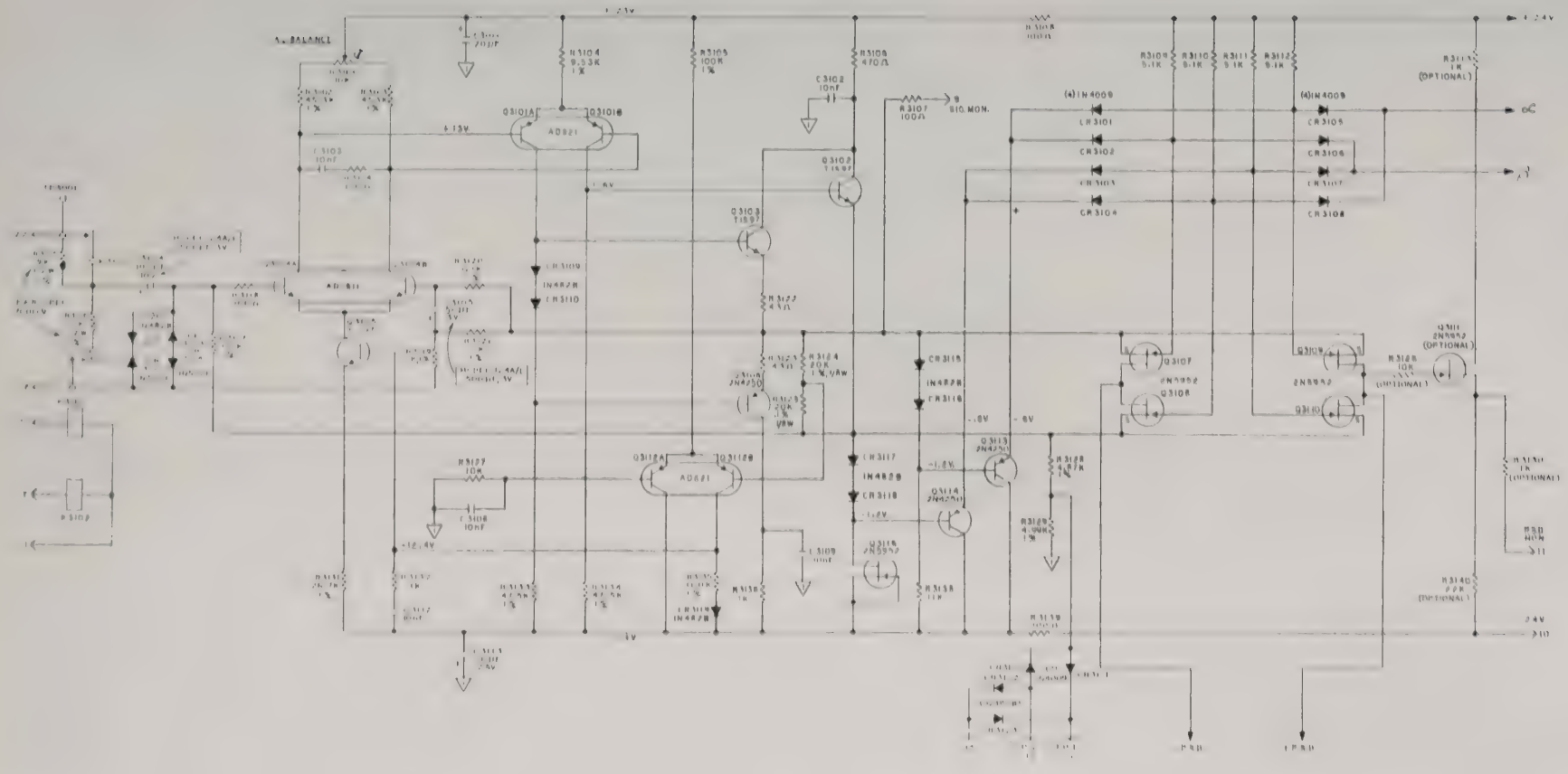
SYMBOLIZATION
MODEL 124A
MIXER BOARD
FAB # 7230-MD-



MODEL 124A

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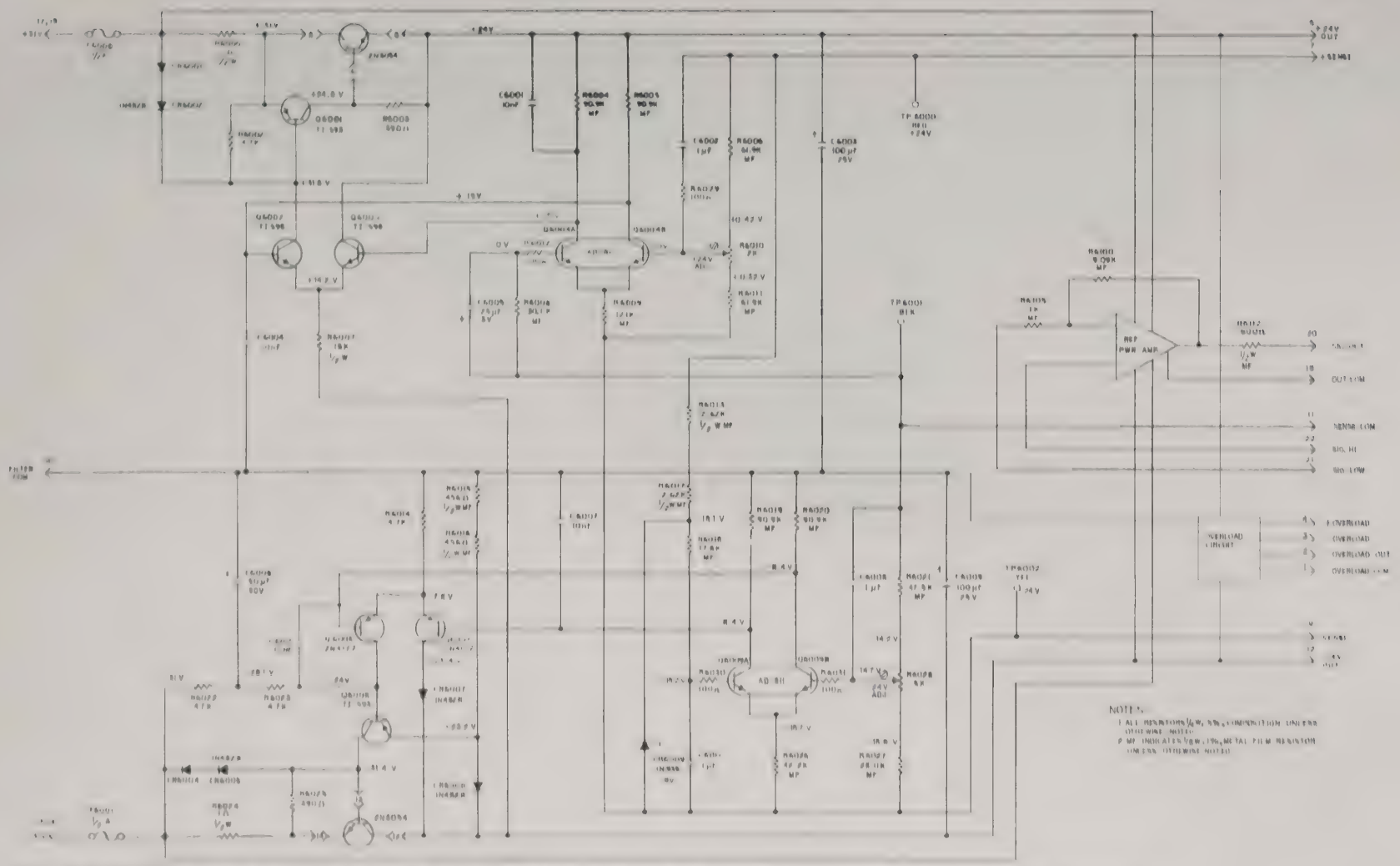

POWER SUPPLY MODEL 124-1-10

SUBASSEMBLY 9240-2000-04 (PARTIAL)

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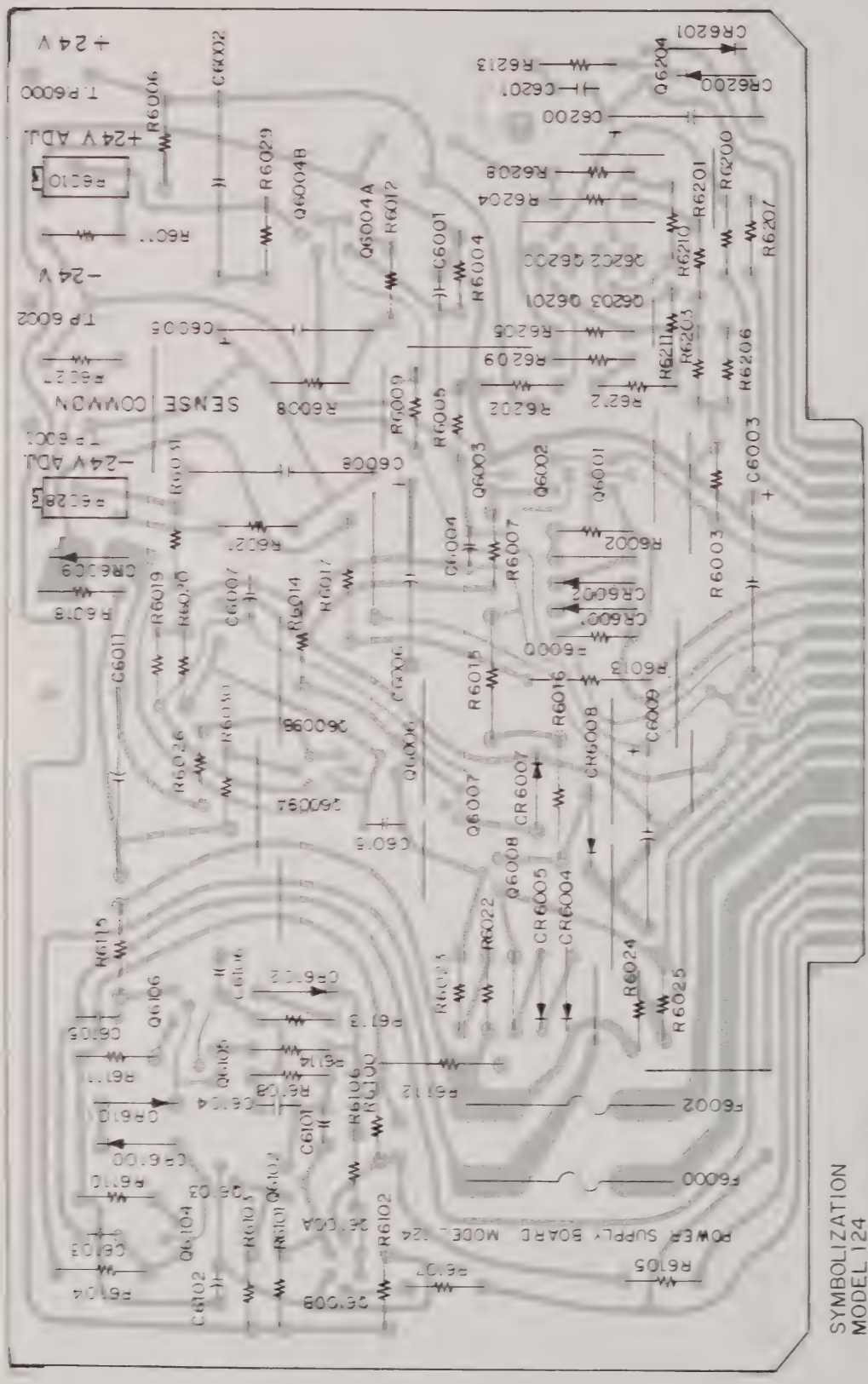
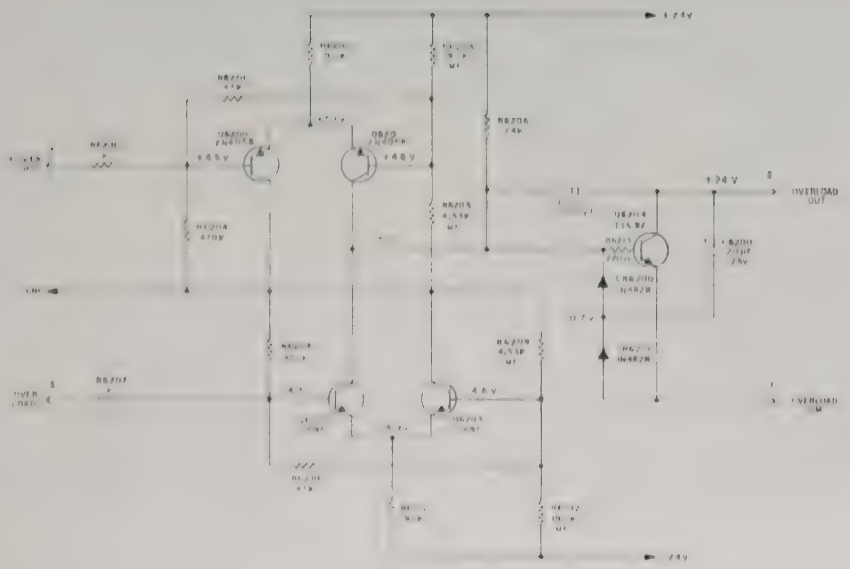
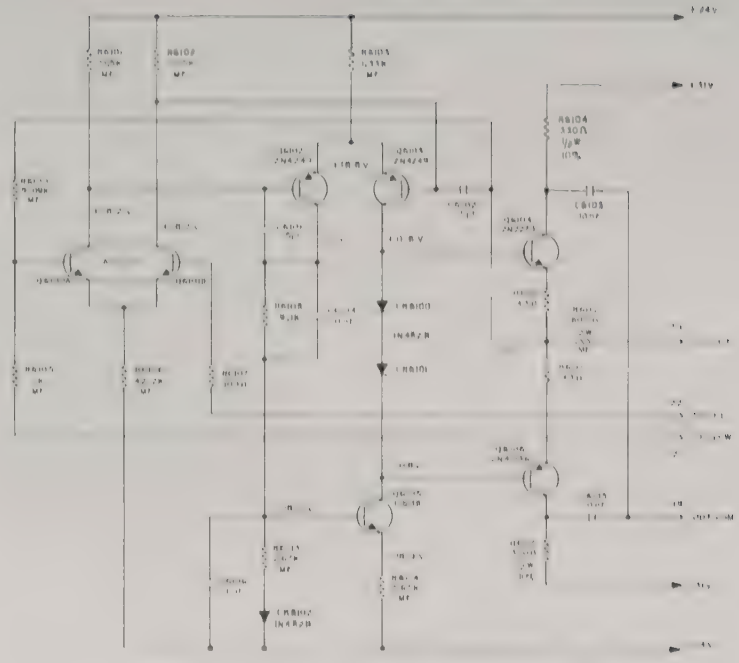


SYMBOLIZATION
MODEL 124
POWER SUPPLY BOARD
FAB # 4986-MD-D



OVERLOAD CIRCUIT



REFERENCE POWER AMP

NOTES:

- ALL RESISTORS ¼ W, 5% COMPOSITION UNLESS OTHERWISE NOTED.
- MF INDICATES ½ W, 1% METAL FILM RESISTORS UNLESS OTHERWISE NOTED.

POWER SUPPLY Model [illegible]

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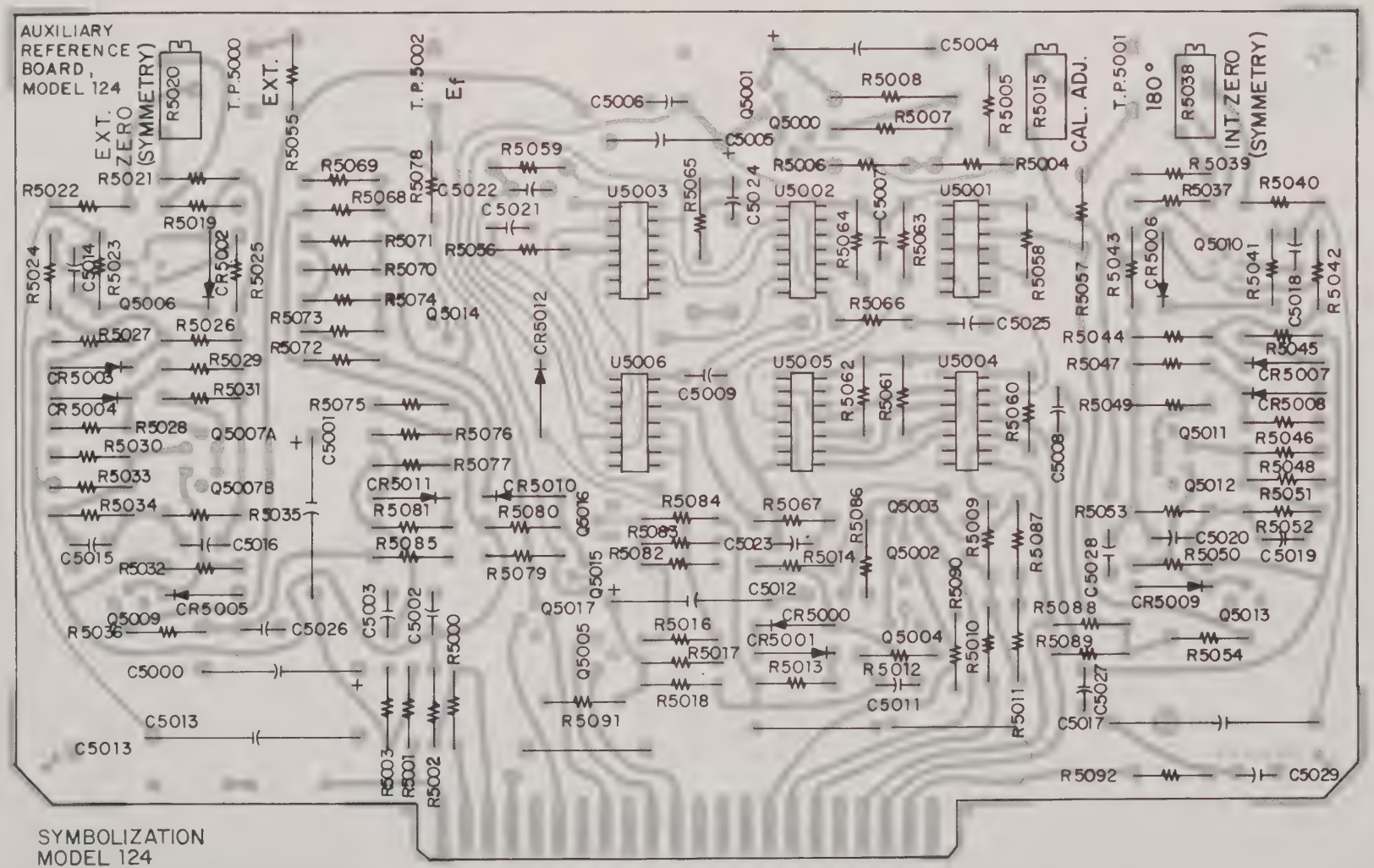
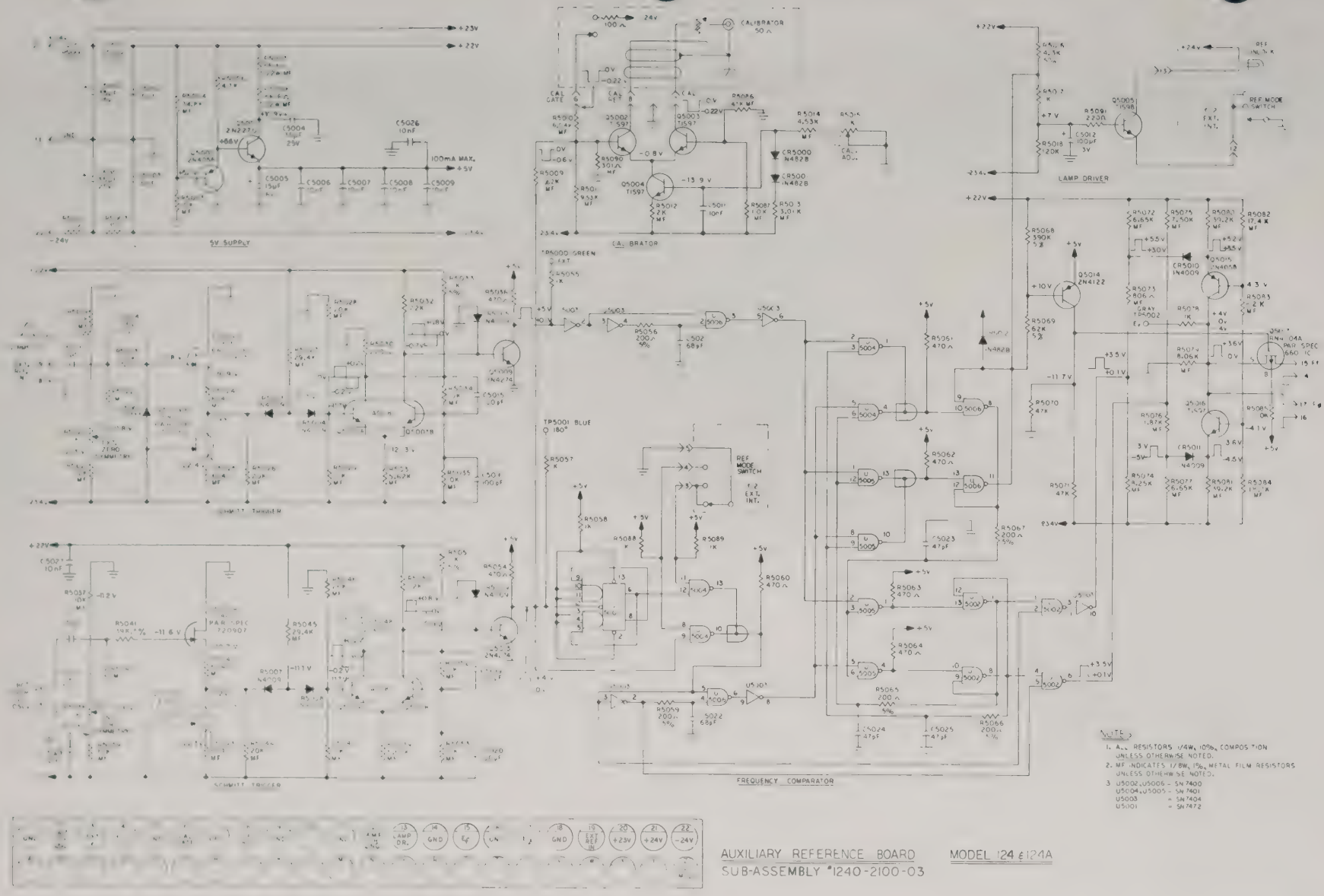
SHEET 2 OF 2 [illegible]



SYMBOLIZATION
MODEL 124
AUXILIARY REFERENCE BOARD
FAB. #4983-MD-D



NOTE

1. ALL RESISTORS 1/4W, 10% COMPOSITION UNLESS OTHERWISE NOTED.
2. MF INDICATES 1/8W, 1% METAL FILM RESISTORS UNLESS OTHERWISE NOTED.
3. U5002,U5006 = SN7400
   U5004,U5005 = SN7401
   U5003 = SN7404
   U5001 = SN7472

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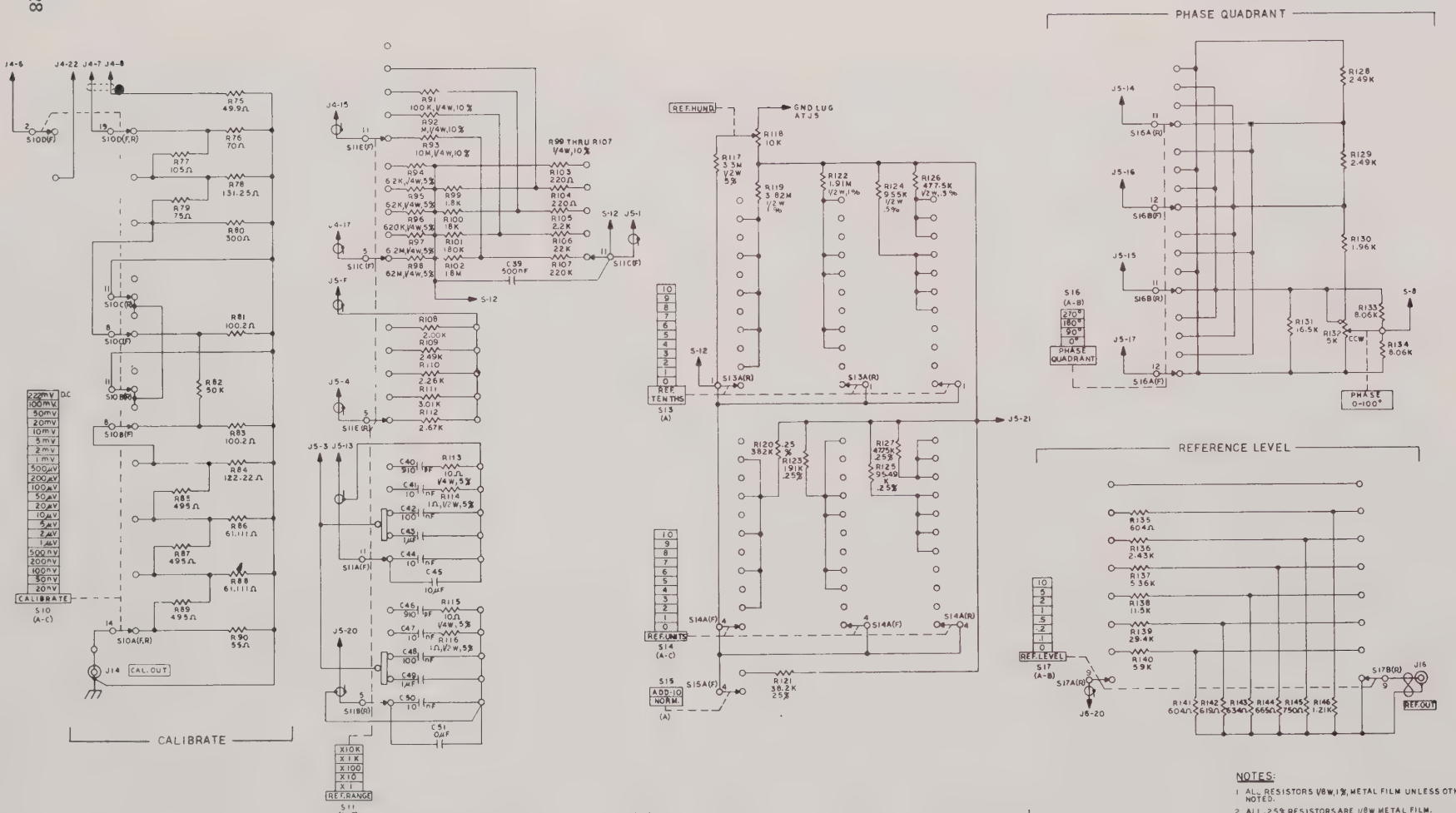
4871 D SD L



SWITCH SCHEMATIC

MODEL 124A

NOTES:

1. ALL RESISTORS 1/8W, 1%, METAL FILM UNLESS OTHERWISE NOTED.
2. ALL .25% RESISTORS ARE 1/8W METAL FILM.
3. ALL 1/4W 5%, 1/4W 10%; 1/2W 5%, 1/2W 10% ARE COMPOSITION RESISTORS.
4. CAPACITORS ON SIGNAL RANGE SWITCH (S3) AND REFERENCE RANGE SWITCH (S11) ARE SET PER PAR. SPEC. #70012.
5. R76 THRU R90 ARE .1%, 1/2W, METAL FILM.

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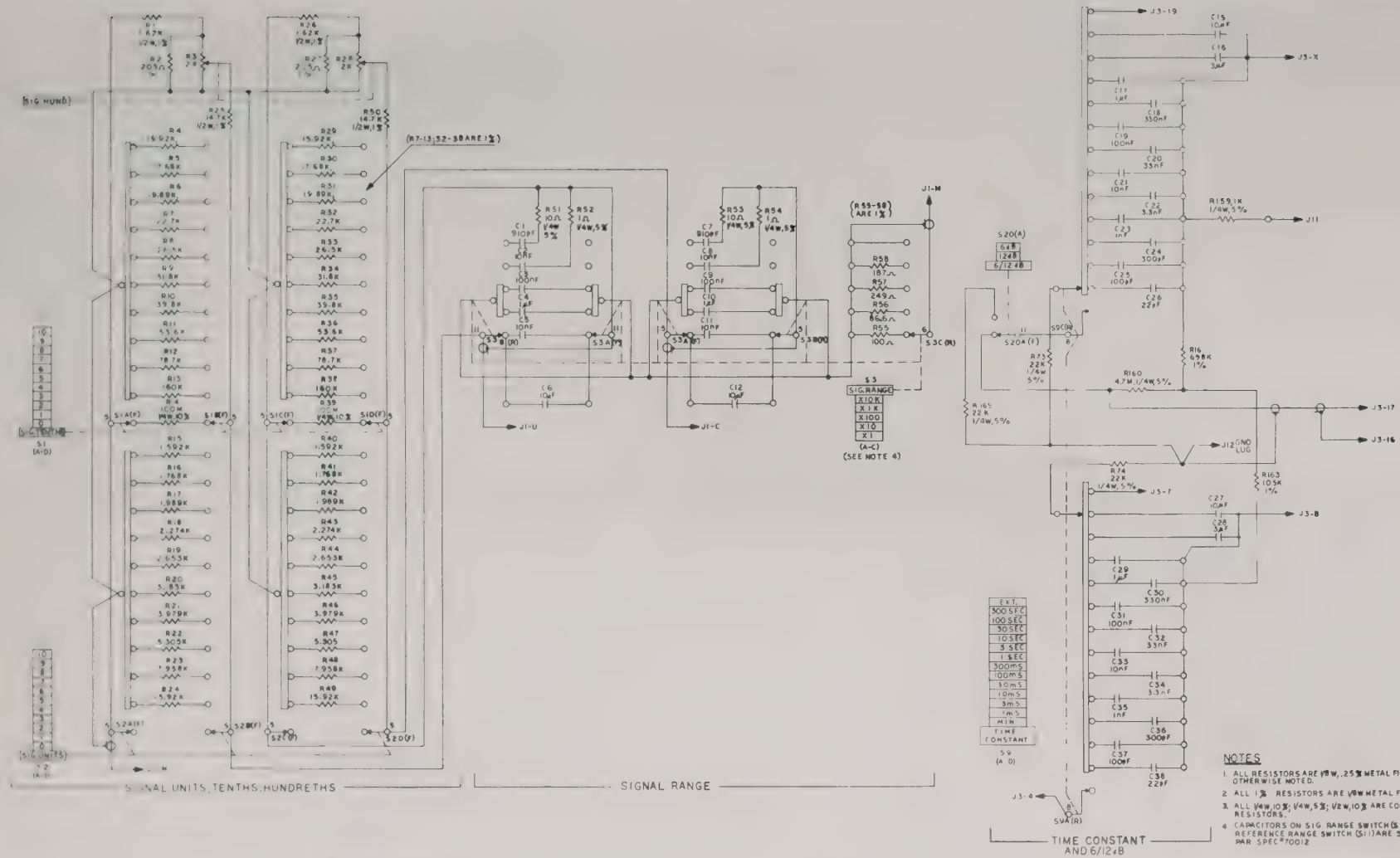

SWITCH SCHEMATIC

MODEL 124A

NOTES

1. ALL RESISTORS ARE 1/8W, .25% METAL FILM UNLESS OTHERWISE NOTED.
2. ALL 1% RESISTORS ARE 1/8W METAL FILM.
3. ALL 1/4W, 10%; 1/4W, 5%; 1/2W, 10% ARE COMPOSITION RESISTORS.
4. CAPACITORS ON SIG RANGE SWITCH (S3) AND REFERENCE RANGE SWITCH (S11) ARE SET PER PAR SPEC #70012

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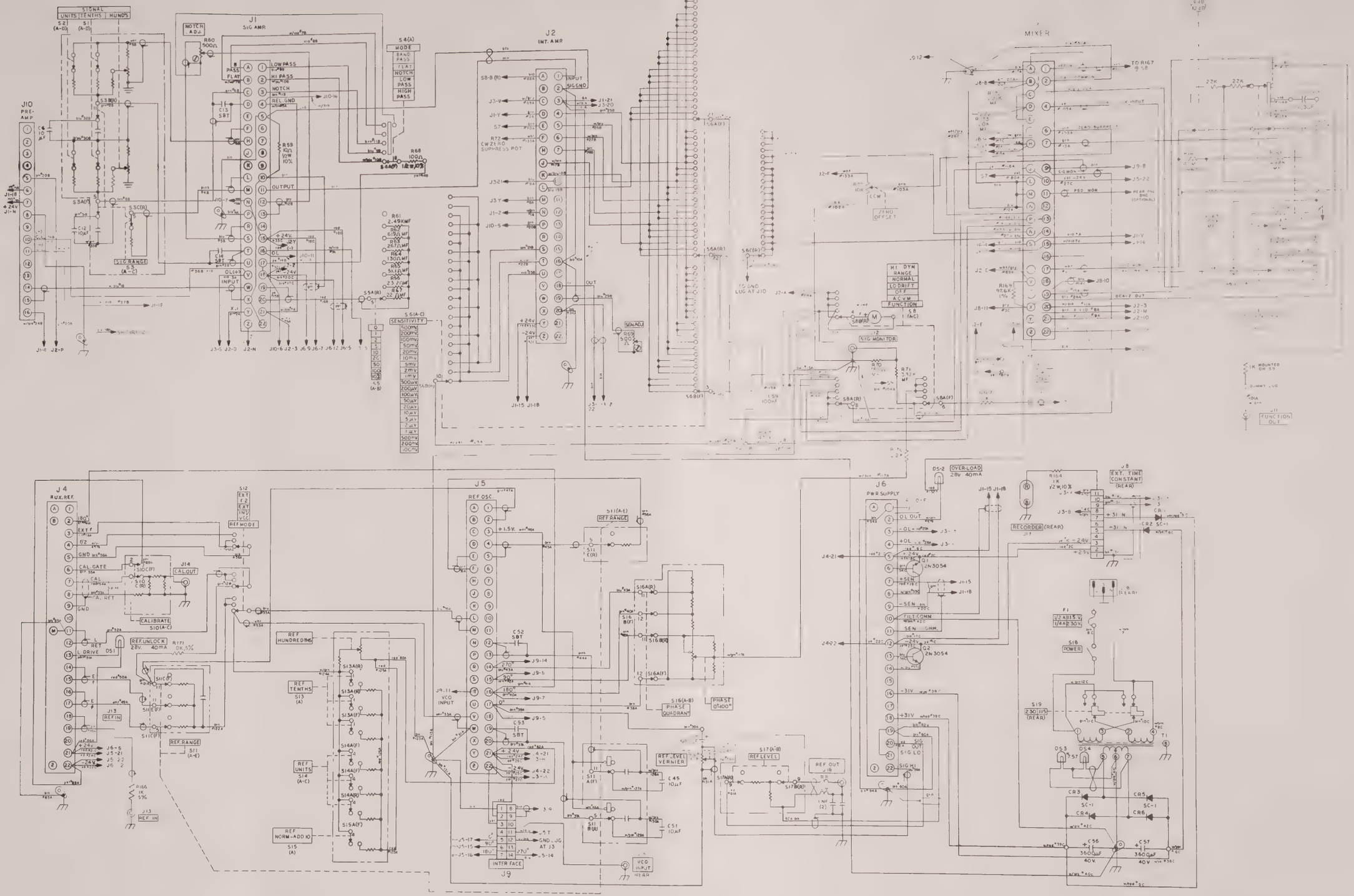

CHASSIS WIRING SCHEMATIC

MODEL 124A

NOTES:

1. ALL RESISTORS 1/4W, 10%, COMPOSITION UNLESS OTHERWISE NOTED.
2. MF INDICATES 1/8W, 1%, METAL FILM RESISTOR UNLESS OTHERWISE NOTED.

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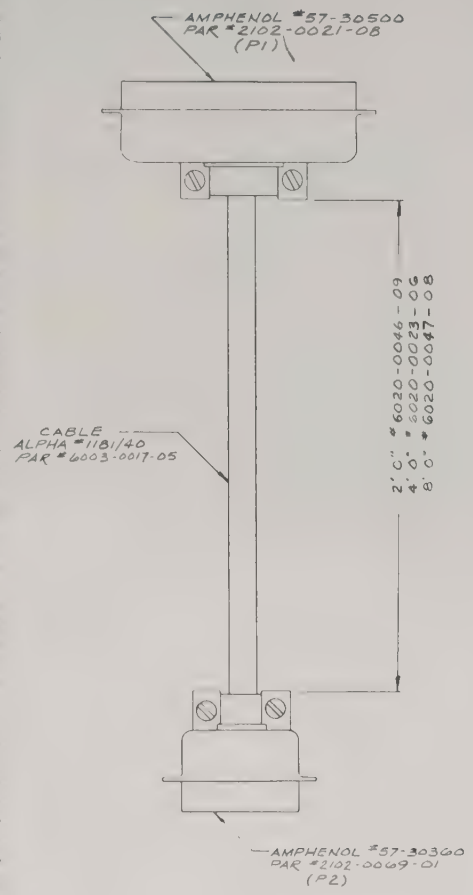
SH 3 OF 3

7720 | E | SD

WIRE				CONNECTOR WIRING	
TYPE	GAUGE	FUNCTION	COLOR	FROM CONNECTOR (P1) #2102 0021 08	TO CONNECTOR (P2) #2102 0069 01
		POLARITY	WHT/BLK/BRN	P1-1	P2-1
		SPARE	WHT/RED/PURP	P1-2	P2-2
		WORD 1, A	WHT/RED/BRN	P1-3	P2-3
		WORD 1, C	WHT/RED/GRN	P1-4	P2-4
		WORD 2, A	WHT/RED	P1-5	P2-5
		WORD 2, C	WHT/RED/BLU	P1-6	P2-6
		WORD 3, A	WHT/BLK/BLU	P1-7	P2-7
		WORD 3, C	TAN	P1-8	P2-8
		WORD 4, A	PINK	P1-9	P2-9
		WORD 4, C	WHITE	P1-10	P2-10
		WORD 5, A	GRAY	P1-11	P2-11
		WORD 5, C	YEL	P1-12	P2-12
		CHID U, A	BRN	P1-13	P2-13
		CHID U, C	PURPLE	P1-14	P2-14
40 COND ALPHA #11811/40	22				
		SPARE	WHT/BRN/BLK	P1-19	P2-19
		SPARE	WHT/RED/RED	P1-20	P2-20
		"	WHT/BLK	P1-21	P2-21
		"	WHT/GRN	P1-22	P2-22
		BUSY	WHT/YEL	P1-23	P2-23
		WORD 6, A	WHT/BLU	P1-24	P2-24
		WORD 6, C	WHT/ORG	P1-25	P2-25
		WORD 1, B	WHT/BLK/YEL	P1-28	P2-28
		WORD 1, D	RED/BLK	P1-29	P2-29
		WORD 2, B	BLK	P1-30	P2-30
		WORD 2, D	[illegible]	P1-31	P2-31
		WORD 3, B	[illegible]	P1-32	P2-32
		WORD 3, D	[illegible]	P1-33	P2-33
		WORD 4, B	[illegible]	P1-34	P2-34
		WORD 4, D	[illegible]	P1-35	P2-35
		WORD [illegible]	[illegible]	P1-36	P2-36
		WORD [illegible]	[illegible]	P1-37	P2-37
		CHID U, B	WHT/BRN	P1-38	P2-6
		CHID U, D	WHT/BLK/BLK	P1-39	P2-10
		[illegible]	[illegible]	P1-48	P2-18
		[illegible]	[illegible]	P1-47	P2-[illegible]
		[illegible]	[illegible]	P1-50	P2-17



REVISIONS				
SYM	ZONE	DESCRIPTION	DATE	APPD.

ITEM	QUAN	PART NO.	DESCRIPTION	DWG. NO

TOL'S. UNLESS NOTED NON-ACCUMULATIVE

FRAC.	DEC.	ANG.
± 1/64	± .005	± 1/2

SCALE NONE

DRWN

CHKD

APPD

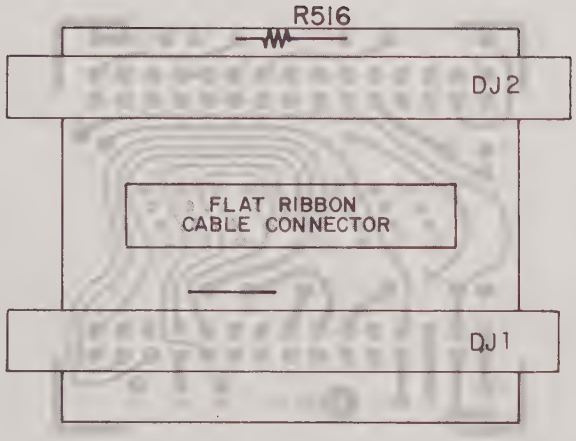
TITLE
CABLE INTERFACE
PAR 262 TO 264,267

USED ON MODELS
262, 264, 2[illegible]

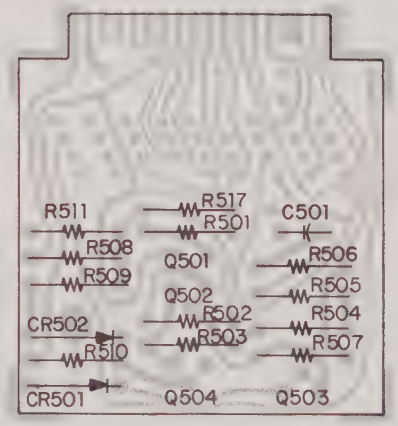
PART NO. | NEXT ASSY

PRINCETON APPLIED RESEARCH
PRINCETON N.J

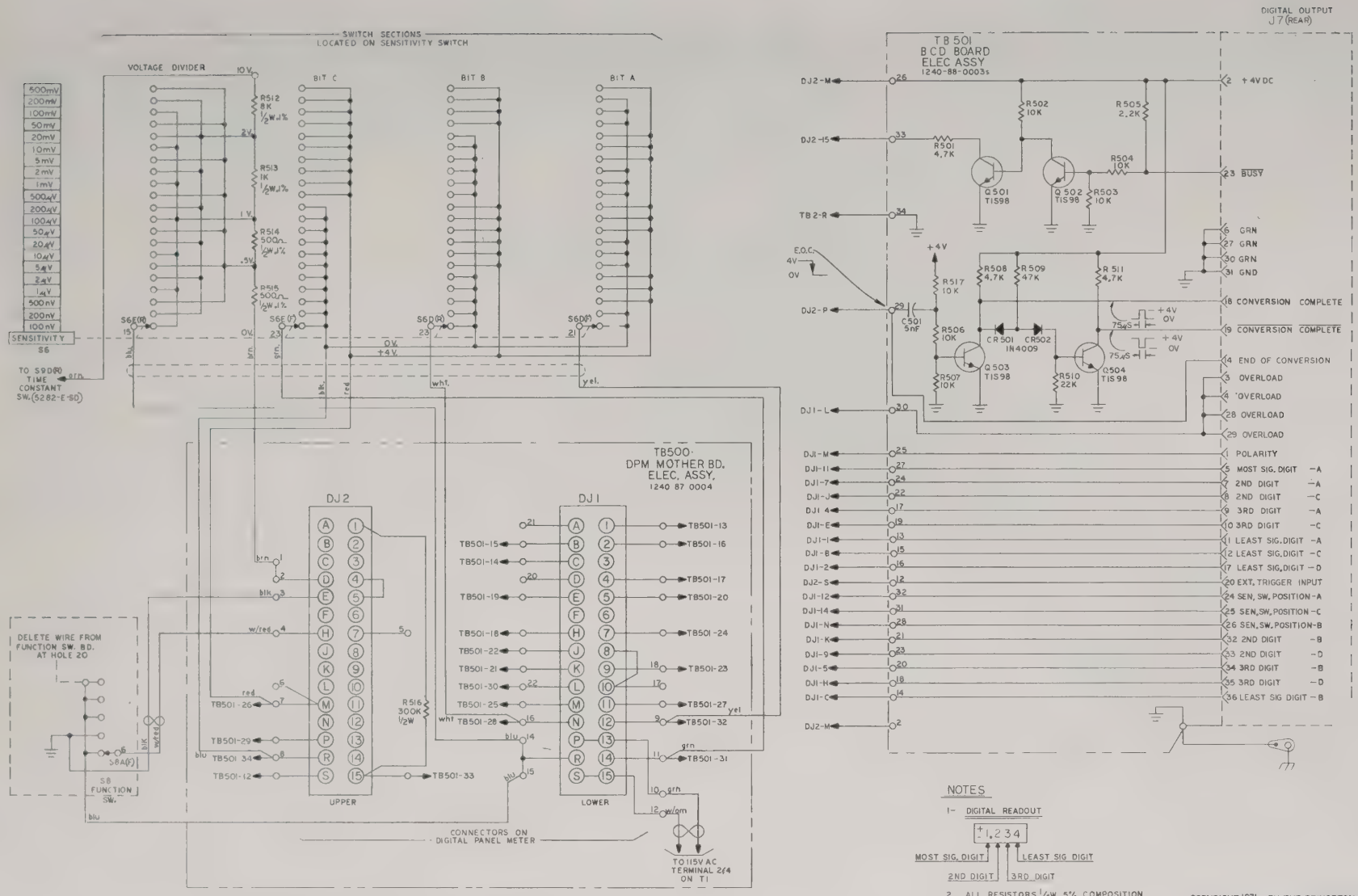
DWG NO	SIZE	TYPE	REV
[illegible]	C	[illegible]	



MODEL 124, 127, 156
DPM MOTHER BOARD



MODELS 124, 127, 156
DVM OPTION BOARD



NOTES

1- DIGITAL READOUT

±1,234

MOST SIG. DIGIT — LEAST SIG DIGIT

2ND DIGIT — 3RD DIGIT

2 ALL RESISTORS 1/4W, 5%, COMPOSITION UNLESS OTHERWISE NOTED.

3- ALL 1% RESISTORS 1/2W, METAL FILM

CHASSIS WIRING DIAGRAM DIGITAL PANEL METER WITH BCD OUTPUT

MODEL 124/98 (SERIAL NO. 476 & UP)

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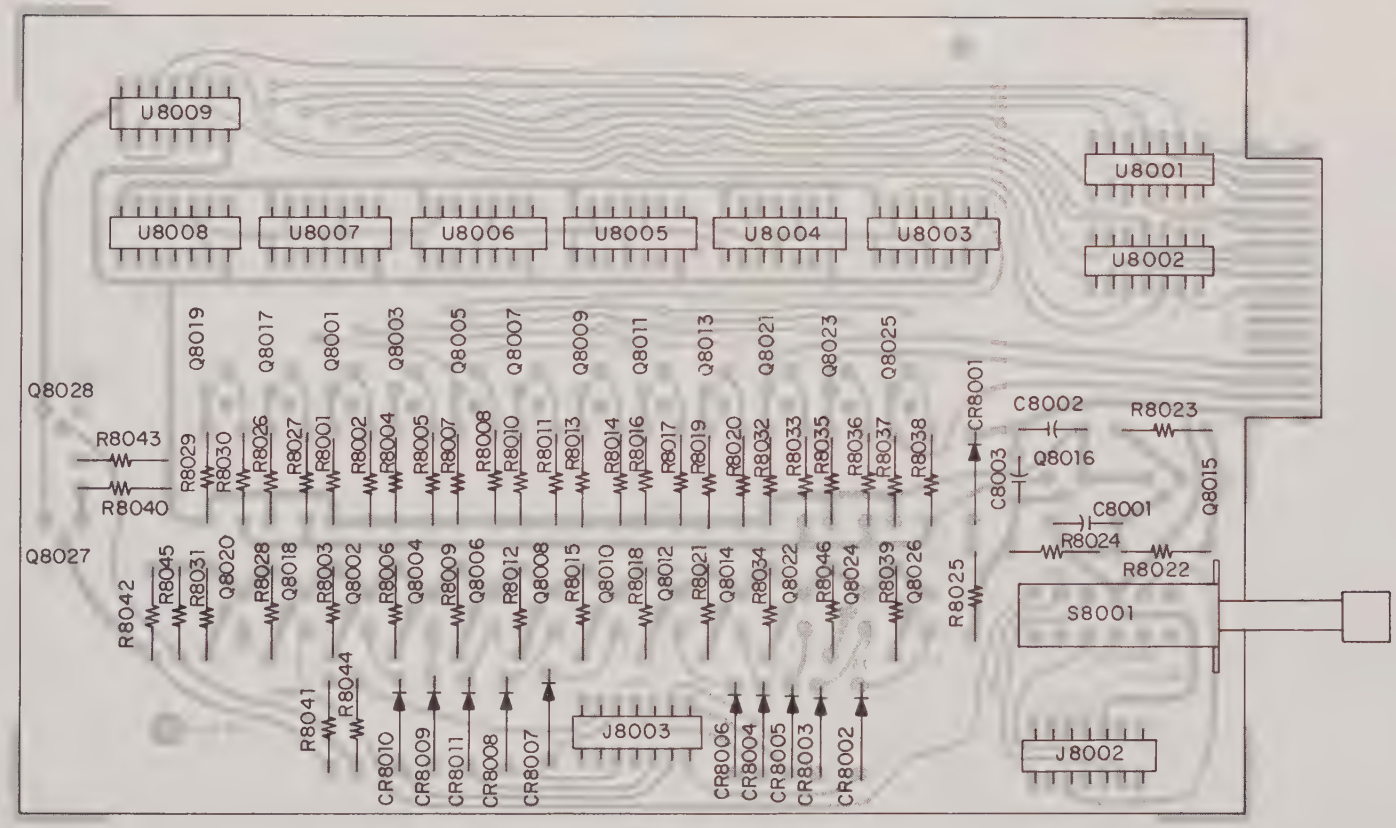
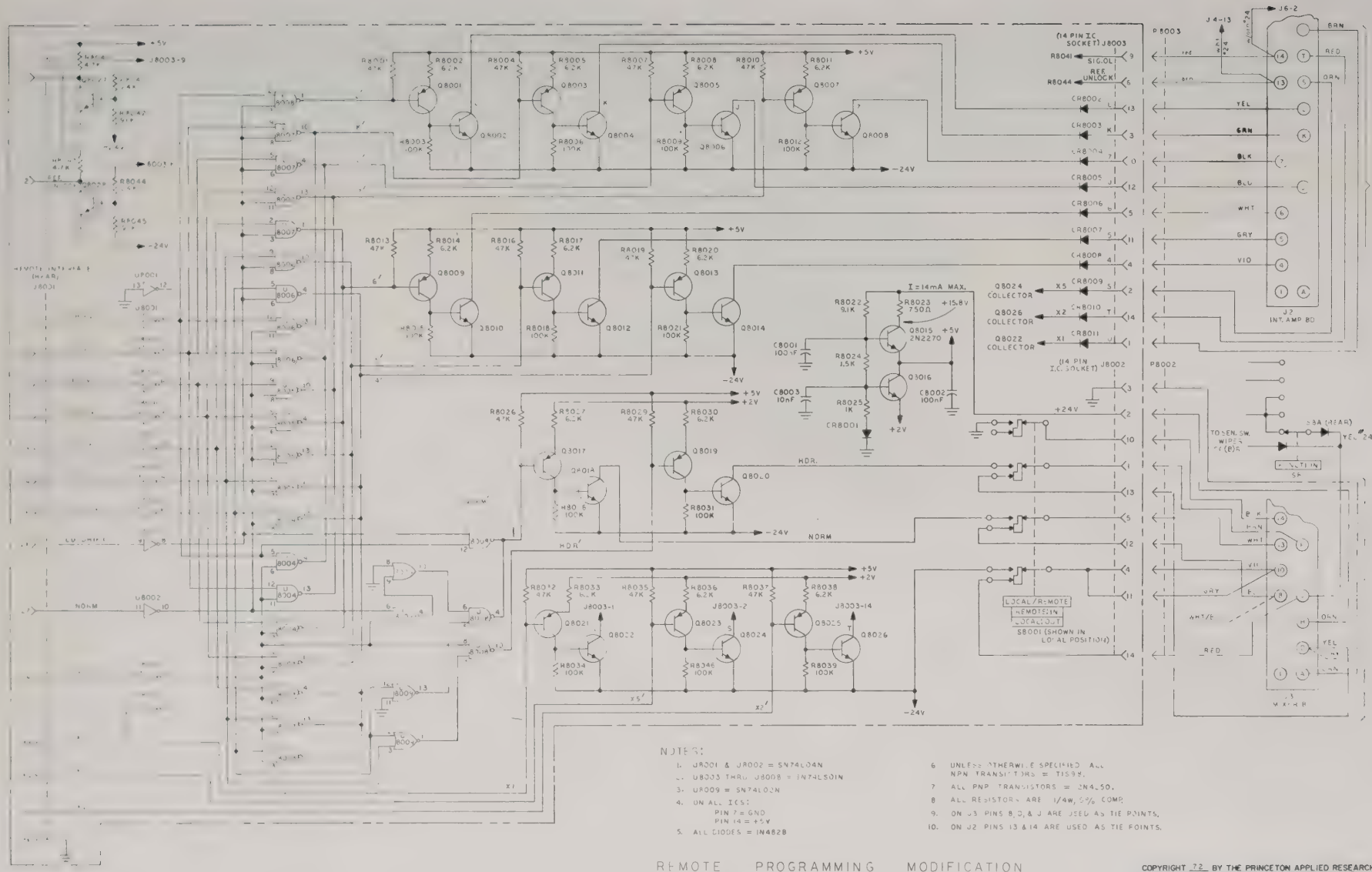
6725 D SD C



SYMBOLIZATION
MODEL 1241/83
REMOTE PROGRAMMING BOARD
FAB. # 7862-MD-A



NOTES:

1. U8001 & U8002 = SN74L04N
2. U8003 THRU U8008 = SN74LS04N
3. U8009 = SN74L02N
4. ON ALL ICS:
   PIN 7 = GND
   PIN 14 = +5V
5. ALL DIODES = IN4828
6. UNLESS OTHERWISE SPECIFIED ALL NPN TRANSISTORS = TIS98.
7. ALL PNP TRANSISTORS = 2N4250.
8. ALL RESISTORS ARE 1/4W, 5% COMP.
9. ON J3 PINS 8, D, & J ARE USED AS TIE POINTS.
10. ON J2 PINS 13 & 14 ARE USED AS TIE POINTS.

REMOTE PROGRAMMING MODIFICATION
MODEL 124A/83 (OPTIONAL)

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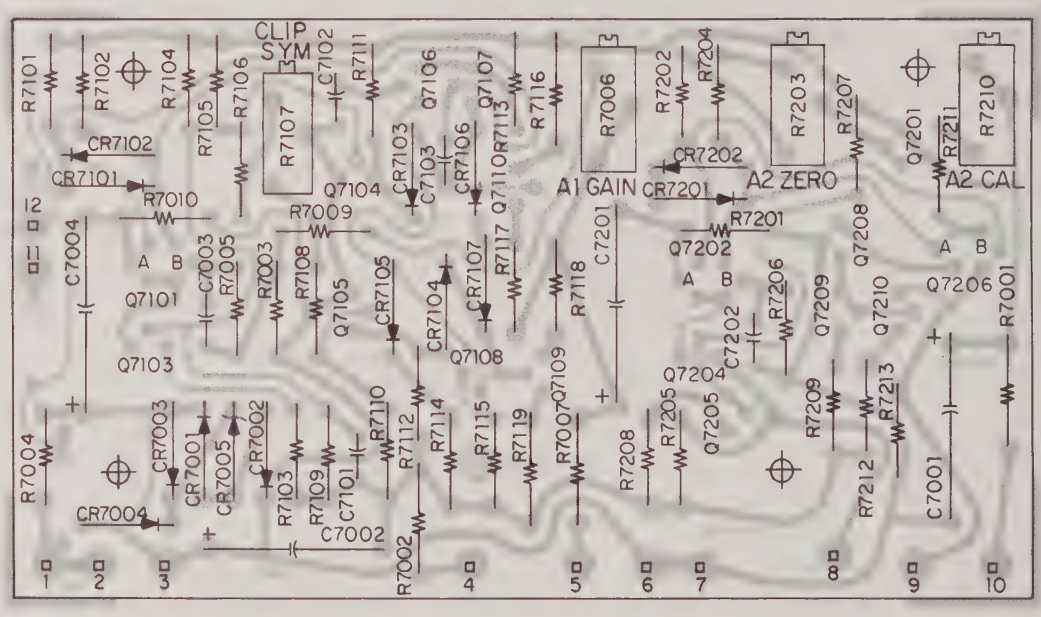
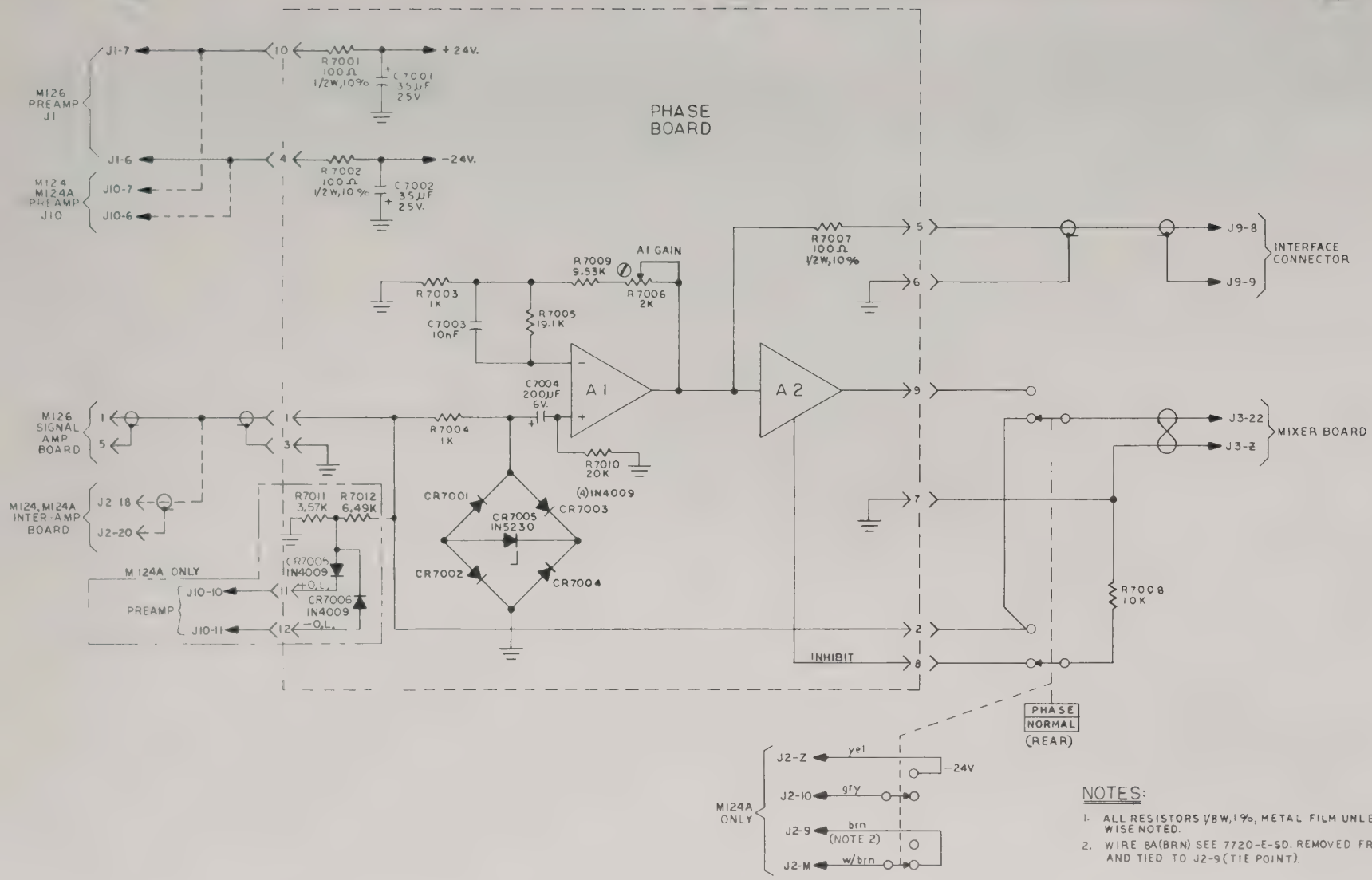
7877 D SD



MODEL 126
PHASE METER OPT. AMPLIFIER BOARD



NOTES:

1. ALL RESISTORS 1/8W, 1%, METAL FILM UNLESS OTHERWISE NOTED.
2. WIRE 8A(BRN) SEE 7720-E-SD. REMOVED FROM J2-M AND TIED TO J2-9(TIE POINT).

PHASE METER MODIFICATION
126/99, 1240/85, 1241/85

MODEL 126
124, 124A

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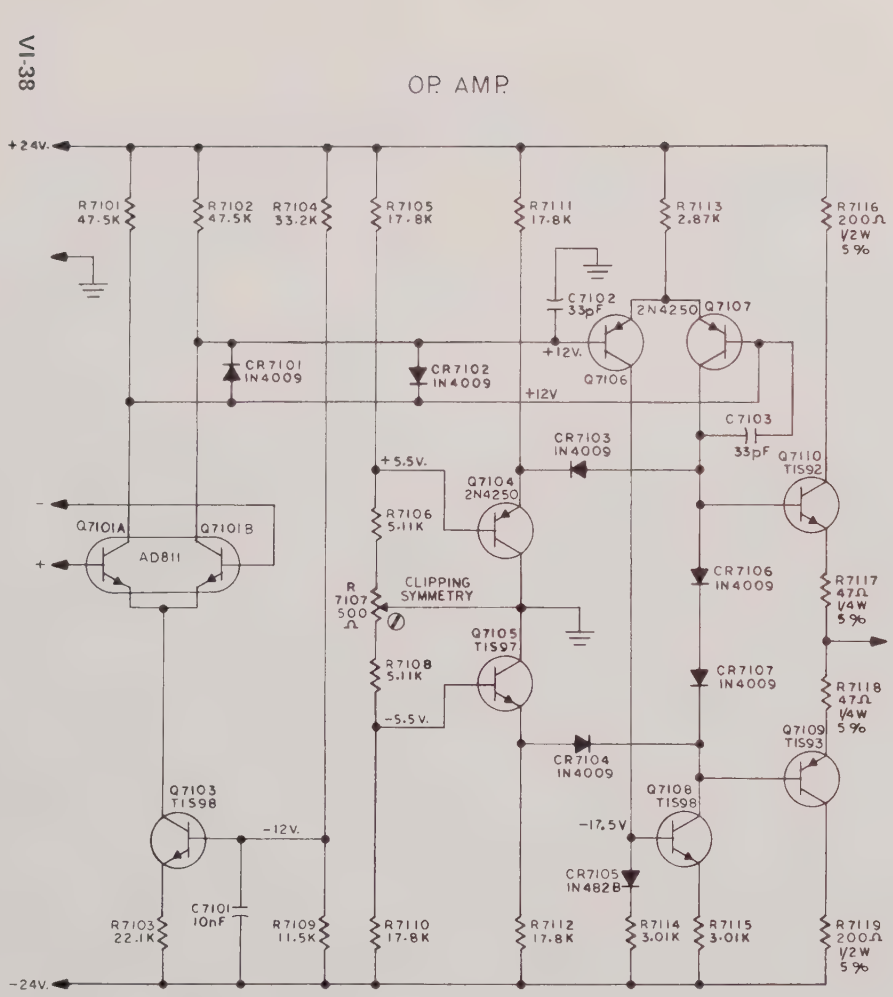
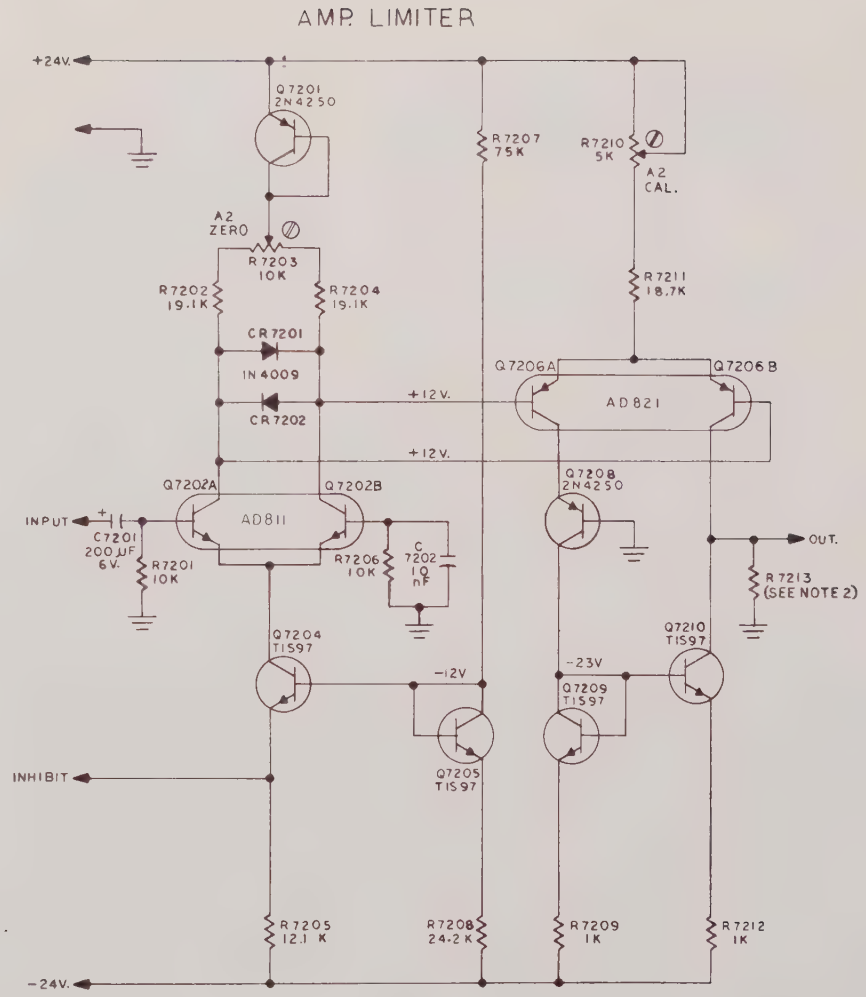
V1-38





AMPLIFIERS

PHASE METER MODIFICATION
1260/99, 1240/85, 1241/85

MODEL 126, 124, 124A

NOTES:

1. ALL RESISTORS 1/8W, 1%, METAL FILM UNLESS OTHERWISE NOTED.
2. MODEL 126 = 15Ω
   MODEL 124 = 150Ω
   MODEL 124A = 75Ω

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APPENDIX A

DYNAMIC RANGE AND THE COMPLETE LOCK-IN AMPLIFIER

CONCEPTS DEFINED

In any signal-processing system, the lock-in amplifier included, there are three critical signal levels that determine the suitability of the system for use in a given application. They are: (1) the FS (Full-Scale Input Level) of the system, (2) the MDS (Minimum Discernible Signal), and (3) the OVL (Overload) level. Figure 1 illustrates how these three signal levels are defined and how they determine the dynamic range characteristics of two lock-in amplifiers, the first a Type 1 Lock-In Amplifier (one having a flat frequency response ahead of the Phase Sensitive Detector), and the second a Type 2 Lock-In Amplifier (one having passband limiting ahead of the Phase Sensitive Detector).

FS and MDS are defined in exactly the same manner for both types of lock-in amplifiers. A system's FS input level is that input signal which provides some arbitrary amount of output, generally selected for convenience in reading and processing, and generally of the order of the system's maximum output excursion capabilities (referred to the input). MDS is defined as the smallest noise-free signal that can be applied to the input and still be recovered at the output. It is determined primarily by the output drift.

An OVL level signal is one large enough to cause some component of the system to become non-linear. No valid measurements can be made while an overload condition exists. Other than output overload, which applies in the same manner for all lock-in amplifiers, there are two different OVL levels to be considered, each of which applies differently according to the lock-in amplifier type. The first, that required to overload the PSD (Phase Sensitive Detector), is called PSD OVL level. In the case of a Type 1 lock-in amplifier, only the PSD OVL level applies. In the case of a Type 2 lock-in amplifier, however, there is a second higher OVL level that must be considered as well. This higher level, defined as the input required to overload the passband limiting (frequency selective) circuits ahead of the PSD, is called the PRE-PSD OVL level.

These three critical levels, FS, MDS, and OVL, give the basis for expressing OUTPUT DYNAMIC RANGE, PSD DYNAMIC RESERVE, and PSD DYNAMIC RANGE, other concepts useful in specifying the dynamic range characteristics of lock-in amplifiers. In the case of a Type 1 lock-in amplifier, PSD DYNAMIC RESERVE and PSD DYNAMIC RANGE are identically the TOTAL DYNAMIC RESERVE and TOTAL DYNAMIC RANGE respectively. However, in the case of a Type 2 lock-in amplifier, they are not. The definitions of these concepts and their bearing on dynamic range performance are explained in the following paragraphs.

Referring to the left-hand illustration of Figure 1, note that the input signal is expressed in multiples of FS. The MDS is indicated as being two decades below FULL SCALE, and the PSD OVL level is indicated as being two decades above FULL SCALE. As indicated in the figure, the OUTPUT DYNAMIC RANGE of the lock-in amplifier is the ratio of its FULL-SCALE INPUT level to its MDS. For the amplifier characterized, the OUTPUT DYNAMIC RANGE is 100:1.

PSD DYNAMIC RESERVE is the ratio of the PSD OVL level to the FULL-SCALE INPUT level, also 100:1 for this amplifier.

The PSD DYNAMIC RANGE (identically the Total Dynam-

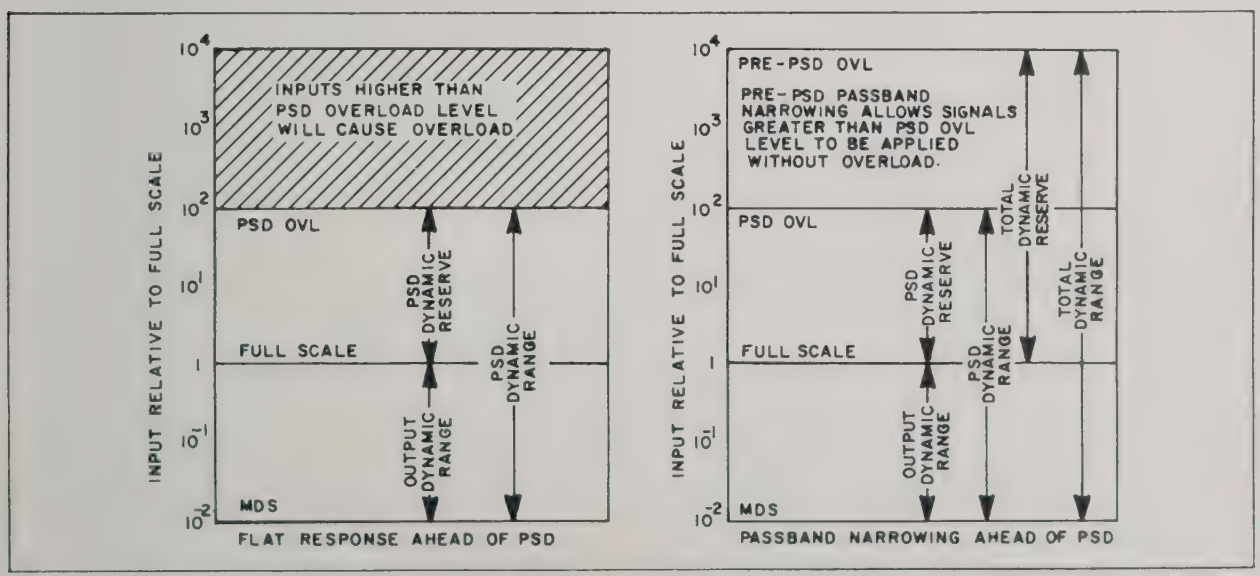

Figure 1. RELATIVE DYNAMIC RANGE CHARACTERISTICS OF TWO LOCK-IN AMPLIFIERS, ONE WITH AND ONE WITHOUT PRE-PSD PASSBAND NARROWING

ic Range for amplifiers having a flat pre-PSD response) is simply the sum (logarithmic) of the OUTPUT DYNAMIC RANGE and the PSD DYNAMIC RESERVE.

What does all this mean in terms of practical applications? First of all, an OUTPUT DYNAMIC RANGE of 100:1 dictates that the smallest signal that can be measured with this lock-in amplifier will be of the order of 10^{-2} of a full-scale input signal. The PSD DYNAMIC RESERVE of 100:1 tells us that interfering signals as high as 100 times full scale can be applied to the input of the lock-in amplifier without overloading it. The PSD DYNAMIC RANGE (10,000:1 in the example) is a kind of "figure-of-merit". In a theoretically perfect lock-in amplifier, the MDS would be small without limit and the PSD OVL level would be large without limit, giving an infinitely large PSD DYNAMIC RANGE. Clearly, all other factors being equal, the larger the PSD DYNAMIC RANGE of a Type 1 lock-in amplifier, the greater the number of applications in which it can be successfully used.

The limitation of a Type 1 lock-in amplifier is simply that the PSD DYNAMIC RANGE is also the TOTAL DYNAMIC RANGE. In other words, inputs greater than the PSD OVL level absolutely cannot be applied without overloading the instrument. With a PSD DYNAMIC RANGE extending over only four decades, the number of applications for the illustrated instrument would be limited.

The right-hand illustration of Figure 1 shows a second lock-in amplifier, one having exactly the same MDS, the same FULL-SCALE INPUT level, and the same PSD OVL level as the first. However, the second amplifier additionally has pre-PSD passband limiting. The characteristics of the pre-PSD signal processing are such as to establish a PRE-PSD OVL level two decades above the amplifier's PSD OVL level. As a result, input signals outside the passband can be as high as 10,000 times FULL SCALE without overloading the instrument. In other words, the TOTAL DYNAMIC RESERVE of this amplifier is 100 times greater than its PSD DYNAMIC RESERVE and so is 100 times greater than the TOTAL DYNAMIC RESERVE (PSD DYNAMIC RESERVE) of the first amplifier as well. Clearly, the second amplifier will always be able to at least

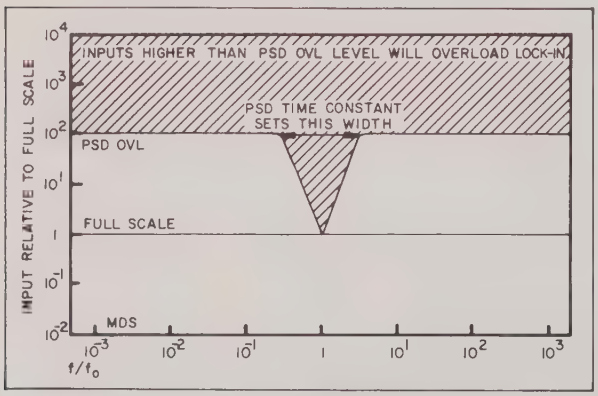

Figure 2. DYNAMIC RANGE CHARACTERISTICS AS A FUNCTION OF f/f_o FOR A LOCK-IN AMPLIFIER NOT HAVING PRE-PSD PASSBAND NARROWING

equal the performance of the first amplifier, and, in applications involving the handling of extremely noisy signals, the second lock-in amplifier can be used to make measurements that would be *completely beyond* the capabilities of the first. All other factors being equal, a Type 2 lock-in amplifier will be useful in a greater number of applications than a Type 1 lock-in amplifier. It is the TOTAL DYNAMIC RANGE that counts, and the amplifier with the greater TOTAL DYNAMIC RANGE will have the greater number of applications. The fact that a given lock-in amplifier without pre-PSD passband limiting may have a little larger PSD DYNAMIC RANGE than another lock-in amplifier with pre-PSD passband limiting is generally of no consequence. It is TOTAL DYNAMIC RANGE that matters.

FREQUENCY CONSIDERATIONS

It may prove instructive to examine the dynamic range characteristics of some lock-in amplifiers as a function of f/f_o, where f_o is the reference frequency and f is the frequency of the applied signal. Figure 2 illustrates the dynamic range characteristics as a function of f/f_o for a Type 1 lock-in amplifier, that is, one not having pre-PSD passband narrowing. As was the case for the Type 1 lock-in amplifier illustrated in Figure 1, this amplifier has an MDS two decades below full scale and a PSD OVL level two decades above full scale. For any lock-in amplifier consisting essentially of a PSD followed by a low-pass filter, the passband is centered on the reference frequency and the passband width varies inversely with the time constant of the low-pass filter. At exactly the reference frequency, the maximum signal one can apply is simply FULL SCALE. On either side of the reference frequency, the maximum tolerable input level rises, reaching a maximum at the PSD OVL level (100 times full scale in the example). By making the passband narrow (the averaging Time Constant large), as much of the interference as possible lies outside the passband, and the likelihood of overload is correspondingly reduced. However, if the input signal outside the passband exceeds 100 times full-scale, absolutely nothing can be done to prevent the overload. **NOTE:** It is implicit in this discussion that one is trying to measure an MDS level signal, which automatically prohibits one from operating with reduced sensitivity to bring the interference below the PSD OVL level. In other words, the Type 1 lock-in amplifier illustrated in the figure absolutely could not be used to measure signals calling for a TOTAL DYNAMIC RANGE greater than 10,000:1.

Now consider the case for a Type 2 lock-in amplifier, that is, one having pre-PSD passband narrowing. The passband ahead of the PSD can be narrowed in several ways. For example, one could use a bandpass filter, a high-pass filter, a low-pass filter, or even a notch filter. Each method has its own advantages and disadvantages, but all have a single common characteristic, namely, that each can be used to extend the TOTAL DYNAMIC RANGE of the amplifier. Because bandpass filtering is probably the most all-around useful technique, it is discussed first.

Figure 3 illustrates the dynamic range characteristics as a function of f/f_o of a Type 2 lock-in amplifier. As before,

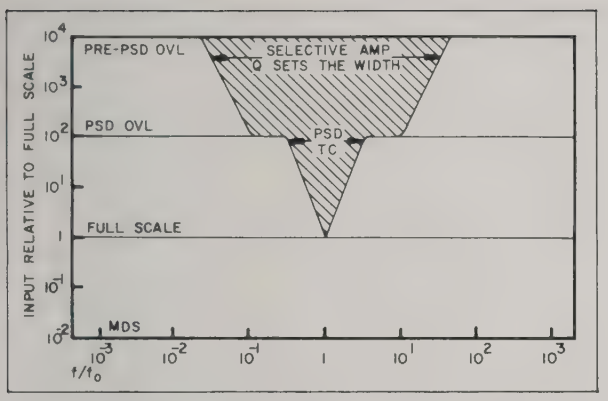

Figure 3. DYNAMIC RANGE CHARACTERISTICS AS A FUNCTION OF f/f_0 FOR A LOCK-IN AMPLIFIER HAVING BANDPASS PRE-PSD PASSBAND NARROWING

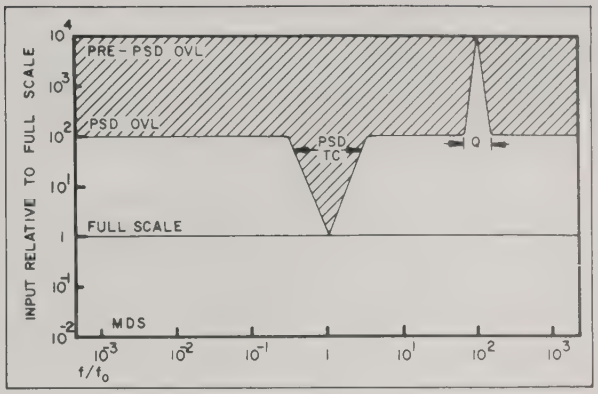

Figure 4. DYNAMIC RANGE CHARACTERISTICS AS A FUNCTION OF f/f_0 FOR A LOCK-IN AMPLIFIER HAVING NOTCH PRE-PSD PASSBAND NARROWING

the MDS, FULL-SCALE, and PSD OVL levels of this amplifier are presumed to be the same as those of the Type 1 lock-in amplifier illustrated in Figure 2. However, the Type 2 lock-in amplifier has a bandpass filter ahead of the PSD that gives a PRE-PSD OVL level two decades above the amplifier's PSD OVL level and, therefore, two decades above the PSD OVL level of the Type 1 lock-in as well.

Consider the overload characteristics of this amplifier as a function of input frequency. As with the first amplifier, the maximum signal that can be applied at exactly the reference frequency is one (FULL SCALE). Above and below the reference frequency, the overload tolerance level rises until the PSD OVL level is reached. Usually, the passband established by the low-pass filter that follows the PSD will be much narrower than that of the pre-PSD bandpass filter. Hence the inverted shoulders shown in Figure 3. Over the frequency range determined by the width of the shoulders, the maximum allowable input is limited to the PSD OVL level. However, as one moves further and further from the reference frequency, the point is reached where the response curves of the bandpass filter are intersected. From these points the maximum acceptable input rises steadily again until the PRE-PSD OVL level is reached. Thus there is a large frequency region over which the illustrated Type 2 lock-in amplifier could accept, without overload, input signals 100 times greater than those the Type 1 lock-in amplifier could accept. The Q of the bandpass filter sets the pre-PSD bandwidth in much the same manner as the time constant of the post-PSD low-pass filter sets the PSD bandwidth. If an experimenter equipped with a Type 1 lock-in amplifier had to measure a signal whose dynamic range characteristics exceeded the PSD DYNAMIC RANGE of the lock-in amplifier, his only possible recourse would be to use other equipment (additional and expensive) ahead of the Type 1 lock-in amplifier to, in effect, convert it to a Type 2 lock-in amplifier. Again, the greater TOTAL DYNAMIC RANGE of the Type 2 lock-in amplifier makes it useful in applications where the Type 1 lock-in amplifier could not be used at all.

Sometimes the interfering signals are concentrated at a single frequency. In applications of this type, inserting a notch filter ahead of the PSD will generally prove to be the best all around solution. Some Type 2 lock-in amplifiers have this capability. Figure 4 shows the same Type 2 lock-in amplifier discussed earlier, but with a notch filter inserted ahead of the PSD instead of a bandpass filter. The width of the notch is set by the filter's Q. Peak rejection is independent of Q, and, for a properly designed filter, will equal the PRE-PSD OVL level.

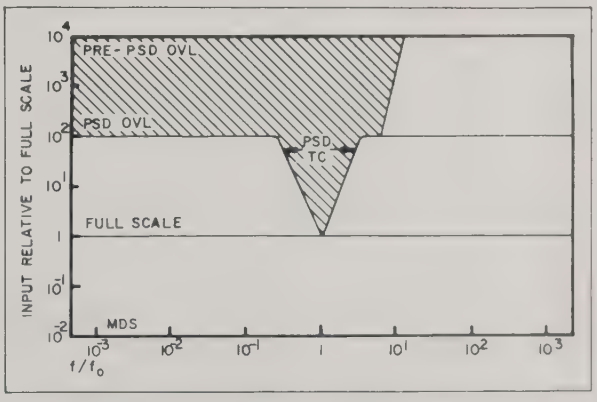

Figure 5. DYNAMIC RANGE CHARACTERISTICS AS A FUNCTION OF f/f_0 FOR A LOCK-IN AMPLIFIER HAVING LOW-PASS PRE-PSD PASSBAND NARROWING

Still other possibilities are to insert a high-pass or low-pass filter ahead of the PSD. These techniques are particularly useful where the interference is concentrated in one band of frequencies while the measurements are to be made by sweeping over another band. Figure 5 shows the low-pass case. The high-pass case would be similar but with the cutoff being to the left of the PSD passband and extending downwards in frequency.

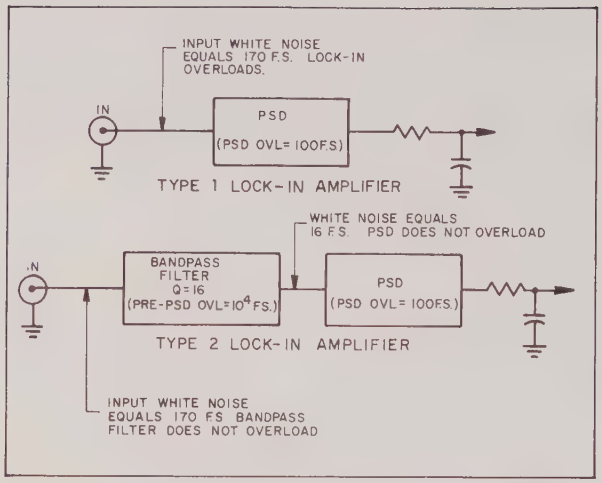
INPUT WHITE NOISE EQUALS 170 F.S. LOCK-IN OVERLOADS.
IN
PSD
(PSD OVL= 100F.S.)
TYPE 1 LOCK-IN AMPLIFIER
WHITE NOISE EQUALS 16 F.S. PSD DOES NOT OVERLOAD
IN
BANDPASS FILTER Q=16
(PRE-PSD OVL=10⁴ F.S.)
PSD
(PSD OVL= 100F.S.)
TYPE 2 LOCK-IN AMPLIFIER
INPUT WHITE NOISE EQUALS 170 FS BANDPASS FILTER DOES NOT OVERLOAD

Figure 6. TYPE 1 AND TYPE 2 LOCK-IN AMPLIFIERS IN APPLICATION INVOLVING HIGH-LEVEL WHITE NOISE AT INPUT

OVERLOAD AND WHITE NOISE

To carry the discussion one step further, consider an example where the interference is broadband white noise. Assume the same lock-in amplifiers are to be tried, that is, a Type 1 lock-in amplifier having a PSD OVL level of 100 times full scale, and a Type 2 lock-in amplifier with the same PSD OVL level but with a PRE-PSD OVL level of 10^4 FS as established by a bandpass filter tuned to the reference frequency. For the example, assume the rms amplitude of the input noise to be 170 times full scale over a 100 kHz bandwidth. Further assume that the reference frequency is 10 kHz. This input noise applied to the Type 1 lock-in amplifier will drive it into overload, rendering it useless for the application at hand.

The Type 2 lock-in amplifier, however, can handle this degree of input noise with ease. The PRE-PSD OVL level is 10^4 FS, sufficient to prevent the bandpass filter from being overloaded. For the measurement to be possible, it is necessary that the noise at the output of the bandpass filter be reduced sufficiently not to overload the PSD, which has an overload threshold of 100 FS. At first glance, one might assume that, if the bandpass filter can reduce the noise to 100 FS, the problem is solved. Unfortunately, even though the nature of white noise makes it necessary to talk about it in terms of rms, real electronic circuits overload on peaks. A generally successful "conversion technique" is to assume that the noise amplitude, from an overload point of view, is about six times its rms value. Thus the noise at the output of the bandpass filter would have to be no greater than 100/6 or about 17 FS. In other words, the noise must be reduced by a factor of ten, from 170 FS at the input of the lock-in amplifier to 17 FS at the output of the bandpass filter. Because the amplitude of white noise varies with the square root of its bandwidth, a factor of ten reduction in amplitude requires a factor of 100 reduction in bandwidth, that is, from 100 kHz at the input to 1 kHz at the output of the bandpass filter. The expression relating the involved parameter is:

$$ENBW = \frac{f_o}{Q} \cdot \frac{\pi}{2}$$

where:

ENBW = Equivalent Noise Bandwidth in Hz
f_o = Tuned frequency in Hz
Q = Tuned frequency divided by 3 dB bandwidth of filter.

By plugging in 1 kHz for the ENBW and 10^4 for f_o, and then solving for Q, one finds that, if the bandpass filter has a Q of 16 or higher, the noise at the output of the bandpass filter will be sufficiently reduced. Again, the Type 2 lock-in amplifier works in a measurement application where the Type 1 lock-in amplifier could not be used at all.

DYNAMIC TRADEOFF

As discussed previously, the TOTAL DYNAMIC RANGE of a lock-in amplifier is divided into two parts, the first being DYNAMIC RESERVE, which describes how much interfering signal outside the passband the amplifier can accept without overload, and the second being MDS, which determines how small a signal can be recovered by the lock-in amplifier. Both are measured relative to FULL SCALE. The TOTAL DYNAMIC RANGE of one lock-in amplifier may be divided differently from the TOTAL DYNAMIC RANGE of another. Furthermore, the fact that the division differs from unit to unit makes each better suited to one type of measurement or another. For example, suppose one had two measurement problems as follows. In the first case, the objective is to measure a full-scale signal having accompanying interference as high as 1000 FS. In the second, the input signal is noiseless, but the lock-in amplifier is to be followed by an analog-to-digital converter having a drift on the order of 10^{-3} FS referred to the input of the lock-in amplifier. Clearly the MDS of the lock-in amplifier must be no worse than 10^{-3} FS if the performance of the system (this includes the analog-to-digital converter) is not to be degraded.

Figure 7 illustrates the dynamic range characteristics of three different lock-in amplifiers, all having a TOTAL DYNAMIC RANGE of 10,000:1 (for the sake of simplicity, only Type 1 lock-in amplifiers are considered). Note that the TOTAL DYNAMIC RANGE is divided differently in each of the three amplifiers. The first has three decades of OUTPUT DYNAMIC RANGE, but only one decade of DYNAMIC RESERVE. The second is divided equally into two decades of OUTPUT DYNAMIC RANGE and two decades of DYNAMIC RESERVE. The third has only one decade of OUTPUT DYNAMIC RANGE, but three decades of DYNAMIC RESERVE. The TOTAL DYNAMIC RANGE of all three lock-in amplifiers spans four decades. Consider the capabilities of each relative to the proposed measurement problems. The first amplifier would not be useful in the high-noise application; it would be hopelessly overloaded. However, in the second application, the one requiring low output drift, the first amplifier would serve nicely. The second amplifier has insufficient DYNAMIC

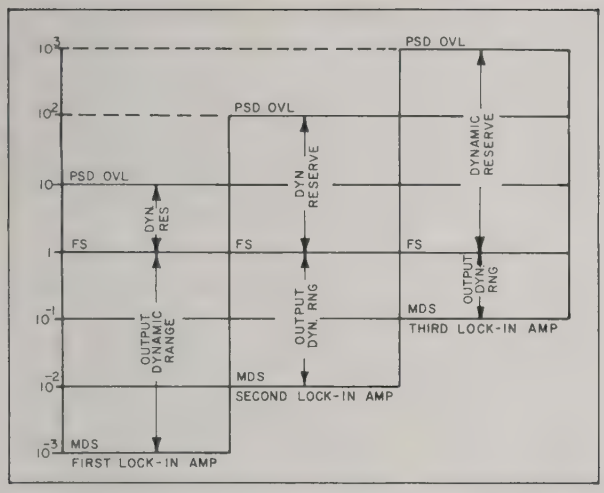

Figure 7. DYNAMIC RANGE CHARACTERISTICS OF THREE LOCK-IN AMPLIFIERS, EACH HAVING THE SAME TOTAL DYNAMIC RANGE BUT WITH A DIFFERENT DYNAMIC TRADEOFF

RESERVE for the first application and insufficient OUTPUT DYNAMIC RANGE for the second. The third amplifier has enough DYNAMIC RESERVE for the first application, but its output drift is 100 times too large for the second. Clearly, in considering the general usefulness of a lock-in amplifier, one has to consider not only the TOTAL DYNAMIC RANGE of the instrument, but also how that TOTAL DYNAMIC RANGE is divided. Ideally, one would like to have a lock-in amplifier for which the dynamic tradeoff could be set to the optimum point for each application, all the while keeping the TOTAL DYNAMIC RANGE constant.

THE COMPLETE LOCK-IN AMPLIFIER

As stated previously, the theoretically perfect lock-in amplifier would be one having (in addition to many other wonderful properties not discussed here) an infinitesimally small MDS level and an infinitely large OVL level. Unfortunately, researchers and experimentalists too live in a real world where nothing is perfect, including lock-in amplifiers. Just the same, it is not out of the question to discuss the real dynamic range characteristics that a lock-in amplifier should have to be good enough for just about every imaginable application. Based on the foregoing discussions, the "complete" (if not perfect) lock-in amplifier should have:

(1) An extremely low MDS, perhaps on the order of 10^{-4} FS.

(2) A very high PSD OVL, perhaps on the order of 10^{3} FS.

(3) Provision for PRE-PSD passband narrowing, adequate to extend the TOTAL DYNAMIC RESERVE out to perhaps 10^{5}.

(4) An extremely wide TOTAL DYNAMIC RANGE; 10^{9} should suffice nicely.

(5) Provision for operator control of the Dynamic Tradeoff so that the dynamic range characteristics of the lock-in amplifier can be optimized for any given application.

IS IT AVAILABLE?

A lock-in amplifier having these dynamic range characteristics can certainly be considered to be "complete", at least with regard to dynamic range capabilities. Moreover, it can be stated that a lock-in amplifier having these dynamic range characteristics is by no means a dream, that it is already available, and that it is the Model 124A.

As shown in Figure 8, the dynamic tradeoff of the Model 124A can be set to three different points, LO DRIFT, NORMAL, and HI DYNAMIC RESERVE, establishing three different sets of dynamic range characteristics. Each dynamic-tradeoff transfer exchanges one decade of OUTPUT DYNAMIC RANGE for one decade of PSD DYNAMIC RESERVE. The PSD DYNAMIC RANGE spans six decades for all possible dynamic-tradeoff options. In addition, the PRE-PSD passband limiting capabilities of the Model 124A allow the TOTAL DYNAMIC RANGE to be extended to nine decades in LO DRIFT and NORMAL operation, and to eight decades in HIGH DYNAMIC RESERVE operation. In addition to Flat (PSD DYNAMIC RANGE limited) operation, Bandpass, Notch, Low-Pass, and High-Pass pre-PSD processing are all offered, each with the Q variable from 1-to-100. Given the Model 124A's wide total dynamic range and dynamic range versatility, it is difficult to imagine a real measurement problem having dynamic range characteristics so severe as to be beyond the capabilities of this instrument.

CAN IT REALLY BE DONE?

Even though measurements under extreme dynamic range conditions are perfectly possible from a theoretical point of view, there are some practical considerations that have to be evaluated carefully before one attempts such a measurement. By way of illustration, consider the following example. Suppose one had a 1000 Hz signal with an amplitude of 10^{-1} FS and accompanied by broadband noise of 10^{4} FS rms over a bandwidth of 160 kHz.

There are three questions that must be considered before proceeding. They are:

(1) Can the pre-PSD passband narrowing reduce the noise enough so that the PSD won't overload?

(2) Can the post-PSD filtering reduce the noise enough so that the output won't overload?

(3) Can the post-PSD filtering reduce the noise enough to allow a reading of the signal level to be taken?

From Figure 8, the PSD overload limit in Hi Dynamic Reserve operation is 10^{3} FS. However, because of the nature of white noise, and because real circuits overload on peaks instead of rms, a more realistic objective for the pre-PSD passband limiting would be to get the noise down to 10^{2} FS rms if peak overloading is to be prevented. If the

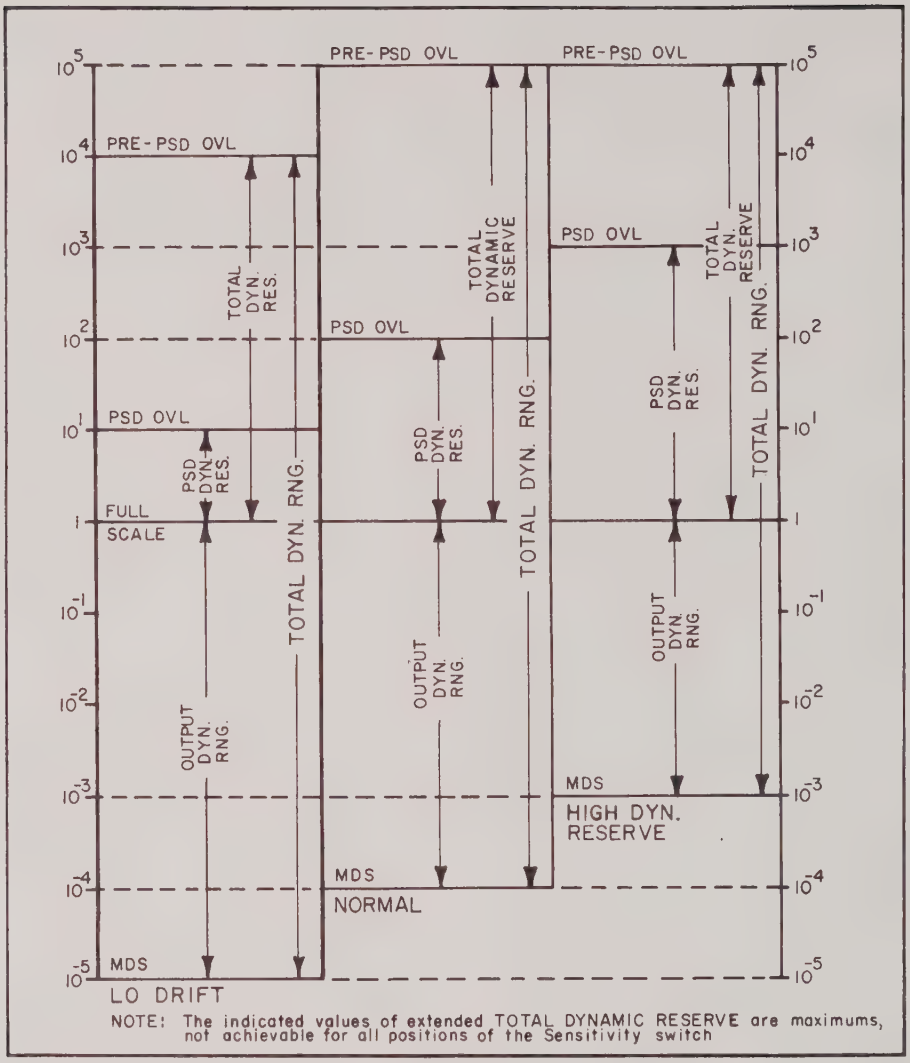

Figure 8. DYNAMIC RANGE CHARACTERISTICS OF THE MODEL 124A

lock-in amplifier is operated with pre-PSD passband filtering, this goal should be readily achievable. Because the noise reduction factor varies with the square root of the bandwidth ratio, a reduction in noise amplitude from 10^4 FS to 10^2 FS will require that the bandwidth be reduced by a factor of 10^4, from 160 kHz to 16 Hz. The expression relating the ENBW and Q of a filter of the type used in the Model 124A is:

$$Q = \frac{f_O}{ENBW} \cdot \frac{\pi}{2}$$

where:

F_O is the tuned frequency and ENBW is the equivalent noise bandwidth, which, we have determined, must be 16 Hz.

By plugging in 10^3 for f_O and 16 for ENBW, we find that the required Q is 100. A Q of 100 is provided in the Model 124A, making the answer to the first question posed above an unreserved "Yes".

A lock-in amplifier overloads at the output if the noise reaching its output exceeds full scale. Again, because of the nature of white noise and real circuits, one might reasonably try to achieve a factor of 10 further improvement, that is, to get the output noise down to 10^{-1} FS to be reasonably confident that output overload on noise peaks will not be a problem. Because the bandwidth achieved by the post-PSD averaging is much narrower than that achieved by the pre-PSD bandpass filtering, the post-PSD bandwidth dominates and can be used solely in determining the output bandwidth. Thus, with regard to question two, the objective is to reduce the noise from 10^4 at the input to 10^{-1} at the output, that is, by a factor of 10^5. This level of noise reduction would require a bandwidth reduction of 10^{10}, from 160 kHz to 16 x 10^{-6} Hz. The expression relating bandwidth and time constant is:

$$TC = \frac{1}{8\ ENBW} \text{ (applies for 12 dB/octave filtering)}$$

Plugging in 16 x 10^{-6} for ENBW and solving for TC, one finds that the required time constant is 7800 s. The

maximum time constant provided by the Model 124A is 300 s, a value much too small to achieve the necessary bandwidth reduction. However, by taking advantage of the external time constant capabilities of the Model 124A, external capacitors could be connected that would extend the time constant to 7800 s and thus allow the output noise to be reduced to .1 FS. The answer to the second question then is also "Yes".

As for the third question, recall that the input signal is presumed to have an amplitude of .1 FS. Our output noise as per the previous discussion is .1 FS rms, a level too high to allow an accurate reading. At the very least, the noise should be reduced by another order of magnitude to 10^{-2} FS rms. A reduction of this order would require a factor of 100 further reduction in bandwidth, which, in turn, would require an increase in time constant by a factor of 100. Thus the necessary time constant would be increased from 7800 s to 780,000 s. The answer to question three is scarcely "Yes". A heart-sinking "Maybe" is the best one can do.

Let us summarize.

(1) Assuming the time constant were increased to 7800 s by means of external capacitors, the lock-in amplifier would not be overloaded and the signal would be available at the output and with the signal-to-noise ratio improved by a factor of 10^5 over what it was at the input.

(2) To read the signal level directly would be extremely difficult. Increasing the time constant to 780,000 s would prove impractical, to say the least. Probably the best possibility would be to use no more external capacitance than is necessary to prevent output overloading. The lock-in amplifier could then be followed by an analog-to-digital converter triggered to make conversions at about one time constant intervals. A digital averager could then be used to average the conversion readings and, after a sufficient number of readings had been averaged, the signal level would be indicated quite accurately. Such a system, although perfectly reasonable from an operating point of view, would require a great deal of time to achieve the desired result. To even run an experiment for the required time would be no small achievement.

The real problem in making such measurements lies, not so much with equipment limitations per se, as it does with human time limitations. To make any given improvement factor in signal-to-noise ratio requires that the averaging time increase by the square of the desired improvement factor. In other words, wide-range lock-in amplifiers such as the Model 124A can handle dynamic range situations generally beyond the operating time available to the researchers who use these instruments. The practical limitations of making linear measurement improvements at the expense of squared experimental time can be formidable. With regard to the question posed at the beginning of this section, "CAN IT REALLY BE DONE?", the answer is a guarded "Yes, it can be, but only if extremely long experiment operating times are available".

WHY BOTHER?

Since it is true that impractically long experimental times are required to make full use of the dynamic range capabilities of an extended-range lock-in amplifier, the reader might reasonably wonder why one should bother considering such an instrument at all. From the point of view of interference consisting purely of white noise, the question is certainly a valid one.

However, not all noise is white noise. Researchers are commonly faced with the problem of measuring a signal which, though accompanied by only moderate levels of white noise, is interfered with as well by high-amplitude transients and/or high-level fixed-frequency signals. Where this is the case, an extended range Type 2 lock-in amplifier is generally the only possible means of measuring the signal successfully. The required time constant is reasonable, the measurement time is short, and the PRE-PSD passbanding limiting gives immunity to the interference. A Type 1 lock-in amplifier, even if it has sufficient PSD DYNAMIC RESERVE to handle the wideband interference, would be so hopelessly overloaded by the transients and/or fixed frequency interference as to cause a complete loss in confidence in measurement data taken. Again, we come back to the statement made earlier, for all around versatility, it's TOTAL DYNAMIC RANGE that counts.

APPENDIX B
GROUNDING

INTRODUCTION

The use of proper grounding techniques is an important consideration in any system processing low-level signals. If grounding is improperly done, ground-loop currents can contaminate the signal and make the measurement unnecessarily difficult, or, in an extreme case, even impossible.

Electromagnetic fields at the power frequency and its harmonics can be found in almost every laboratory (as well as everywhere else). They come from a wide variety of sources; motors, transformers, coils, ac power distribution lines, etc. In short, unless one is willing to build a battery powered installation on the moon, or give architects and contractors nightmares, these fields just cannot be avoided. To further complicate the situation, almost every measurement system contains ground loops, that is, closed circuits formed by tying together all system grounds to bring them to the same potential. Although ground loops are not quite so ubiquitous as power frequency fields, getting rid of them altogether is generally a more formidable task than one should have to contend with in making a measurement. Wherever one has shields and ground loops, one has, in effect, a transformer, with the field sources serving as primary windings and the ground loops serving as shorted one-turn secondary windings. In addition to the ground-loop currents that flow as a result of transformer effects, there are ground-loop currents that result when ground at one point in the system is at a markedly different potential than it is at another point in the system. This frequently happens in large systems where different system components may be plugged into different ac distribution systems. In any case, the experienced experimenter knows that wherever he has ground loops, he has power frequency currents flowing in those ground loops. These currents may or may not pose a problem, depending on the specific circumstances.

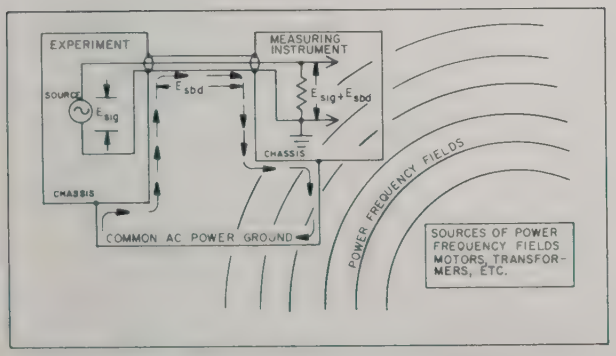

Figure 1. GROUND-LOOP INTERFERENCE

Figure 1 is a block diagram of one possible simple system showing how ground-loop currents arise and how they can interfere with a signal measurement. Referring to the figure, note that there is a ground loop consisting of the common ac ground return line (third wire of the power distribution), the chassis of the experimental apparatus, the braid (shield) of the signal cable, and the chassis of the measuring instrument. Also note that this loop is cut by an alternating electromagnetic field at the power frequency, and that ground-loop currents flow in the loop as a result. The signal to be measured is E_{sig}, defined as the signal produced by the experiment and measured directly across the source. This signal is routed by a shielded cable to the measuring instrument, which "sees" the potential difference between the center conductor and the shield of the cable. As long as this potential difference is identically E_{sig}, only E_{sig} is measured.

However, inasmuch as the cable shield is part of the ground loop, there is a voltage drop, E_{sbd} (sbd = signal-braid drop), from one end of the shield to the other. E_{sbd} equals the product of the shield resistance times the ground loop current. Thus the braid potential at the instrument end of the cable differs from the braid potential at the source end of the cable by E_{sbd}. Because the instrument "sees" the potential difference between the center conductor and the shield (instrument end), the instrument has as its input signal $E_{sig} + E_{sbd}$. In other words, ground-loop signal does indeed interfere with the signal of interest. Whether this interference poses a problem depends on the relative sizes of E_{sig} and E_{sbd}, and on the measurement accuracy required. Unfortunately, in all too many measurement applications, it does pose a problem, and a serious one at that. Some possible solutions are discussed in the following pages.

THE STRAP-THEM-TOGETHER APPROACH

Figure 2 illustrates one popular method of reducing ground-loop interference. This method is based on the premise that, if the resistance between two points is made low enough, the two points will surely be at the same potential no matter what.

In a situation where the ground-loop interference is really not all that bad, and where the resistance is rather high between signal-source common and the input ground of the monitoring instrument, a really significant improvement may be achieved by shorting the chassis together with a ground strap. (Ideally the short should be connected directly from signal-source common to the input ground reference point of the input amplifier.) However, if the interference is very bad, and the resistance is already on the order of a few milliohms, the likelihood of success by this method is rather low. The best one can do is to reduce the ground-loop signal by the ratio of the resistance without the ground strap to the resistance with the ground strap. If

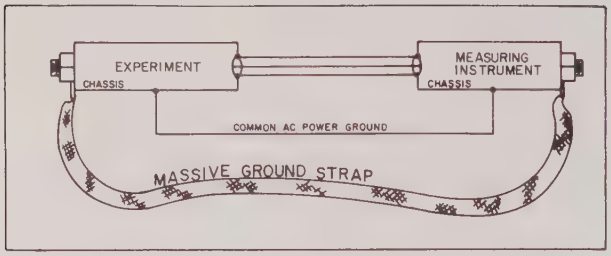

Figure 2. THE STRAP-THEM-TOGETHER APPROACH

the resistance is low to begin with, one soon reaches a practical limit as to how big a ground strap one can use. After all, battery cables are only made so large, and then there's the problem of how to connect massive straps to the points of interest. All in all, the ground-strap method, although useful occasionally, is of limited value.

THE FLOAT-THE-INPUT METHOD

By this method, there is no direct connection between the signal-cable braid at the instrument end of the cable and the chassis of the monitoring instrument. As shown in Figure 3, the cable passes through an opening in the chassis (or an isolating type connector is used) and the braid is connected directly to the ground end of the input resistance. The floating input breaks the ground loop and no ground-loop current flows in the shield. With no ground-loop current flowing in the shield, the potentials at both ends of the cable shield are the same, and the signal developed across the input resistance of the measuring instrument is identically E_{sig}. This method really does work. By using a preamplifier having a floating input, ground-loop interference is eliminated. However, floating-input preamplifiers have other problems. For example, suppose one wished to operate off ground, such as might be the case in measuring a signal developed in a bridge. The preamplifier would have to function properly in the presence of a large common-mode input. Most floating-input preamplifiers would be in deep trouble in this kind of application.

OTHER BREAK-THE-LOOP METHODS

Floating the input is just one example of that broader class of techniques by which the loop is broken so that no ground-loop current flows in the signal-cable braid. In a system larger or more complicated than that illustrated in Figure 3 (and most systems are larger and more complicated), there are many different ground loops, which may or may not share various elements, all of which can contribute to the flow of ground-loop current through the signal-cable braid. Obviously, if the floating input technique can be used, no ground-loop current can flow in the signal-cable braid. One need not care a whit that ground-loop currents continue to flow in other loops as long as their flow through the signal-cable braid is prevented. However, where preamplifier design and construction considerations make it necessary to connect the input cable as depicted in Figure 1, floating input operation is impossible and the operator must seek other means of stopping the ground-loop current interference. In the case shown in Figure 1, the solution would be fairly simple. Either the measuring instrument or the "experimental apparatus" could be floated by using an ordinary ac isolation plug. Nothing more would be required to break the ground loop. However, real systems generally consist of several chassis, all tied together via the third-wire ground return, and often interconnected as well by a veritable jungle of cables carrying all kinds of different signals. Not only that, the various chassis may all be bolted into the same rack, establishing still another path by which they are connected. The difficulties involved in trying to float everything so that all of the loops are broken can get out of hand. Additionally, one should not forget that the reason all the chassis are tied together in the first place is for operator safety in the event of a short circuit. All the more reason for not getting carried away in the pursuit of broken loops. Nevertheless, it could happen that breaking one or two of the worse offenders could result in significant improvement.

THE SEMI-FLOATING GROUND

Figure 4 shows the method used in most of the preamplifiers manufactured by Princeton Applied Research Corporation. The shield of the input cable is grounded to the chassis, but not directly. Instead, some resistance is interposed between the shield and the chassis. In most of these preamplifiers, the shield side of the input floats off ground by ten ohms, although the figure can vary up to as high as 1000 ohms (Model 185). The semifloating ground achieves almost as good a rejection as the true floating input. By inserting the series resistor, the ground-loop

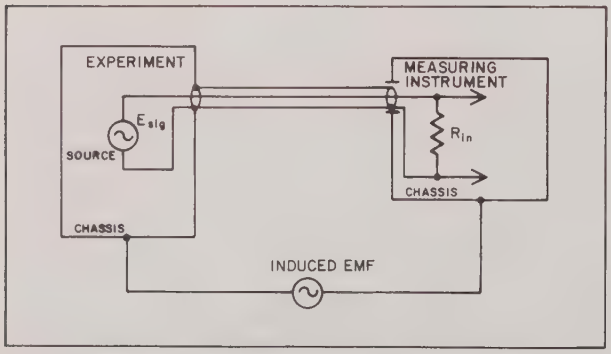

Figure 3. THE FLOAT-THE-INPUT APPROACH

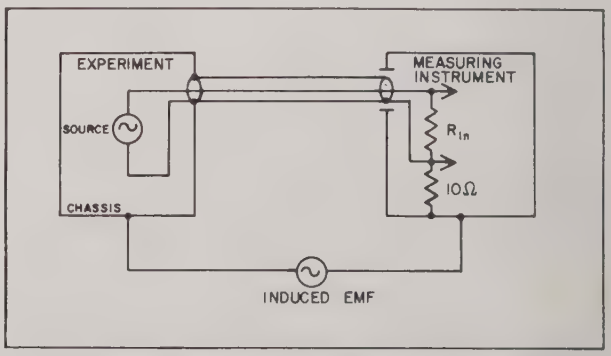

Figure 4. SEMI-FLOATING INPUT GROUND

currents are attenuated and the major part of the ground-loop voltage is developed across the resistor instead of across the braid. In effect, the ground-loop signal is reduced by the ratio of the resistor (typically 10 ohms) to the braid resistance (typically a few milliohms). This method works and works well. Unless the ground-loop signal is really bad, satisfactory single-ended operation with a minimum of trouble is easily achieved.

THE DIFFERENTIAL-FOR-EVERYTHING APPROACH

Strapping-Them-Together, Floating-The-Input, Semi-Floating-The-Input-Ground; each will do the job in certain applications. However, each has its shortcomings. What is needed is a good general approach that is straightforward, always works, and can be used just about everywhere. Well, there is such a method, and that is simply to operate differentially, even when measuring signals from an inherently single-ended source.

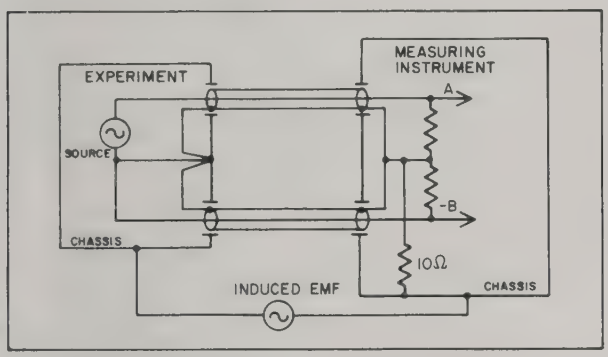

Figure 5. DIFFERENTIAL MEASUREMENT OF "SINGLE-ENDED" SIGNAL

This technique, which is illustrated in Figure 5, should be used whenever possible. All Princeton Applied Research preamplifiers but the special purpose Models 184 and 185 can be used in this way. The Input Selector switch is set to A-B and two input cables are used, with the center conductor of one going to the source and the center conductor of the other going to source common. At the source end, the braid of both cables is returned to source common. At the other end (lock-in amplifier), the floating-ground resistor serves to attenuate the ground loop drop across the braids in the manner discussed earlier. However, whereas in the earlier examples, the amplifier "looked" at the potential difference between the center conductor of the cable and the braid, in this example, the amplifier "sees" the potential difference between the "A" Input and the "B" Input. The ground-loop current flowing in the signal cable braid is of no consequence. Also, the high common-mode rejection characteristic of these preamplifiers assures that common-mode power frequency pickup will not be a problem either.

When operating differentially, there is only one precautionary measure to be tended to, and that is to be sure that any common mode signal arising in ground loops is just that, that is, without a significant differential component. Such a differential component could be developed if the

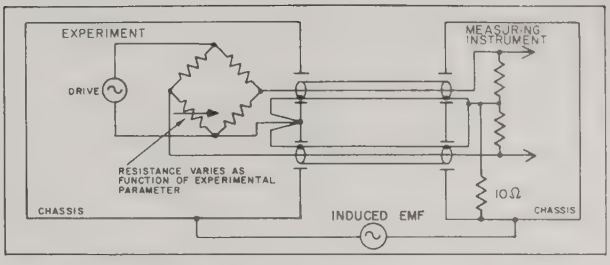

Figure 6. DIFFERENTIAL MEASUREMENT OF "OFF-GROUND" SIGNAL

two input cables followed different paths. However, by using two cables of the same length and following the same path with both, this problem can be avoided.

Figure 6 illustrates the use of differential operation to process an "off-ground" signal. The most important consideration in an application of this type is that the common-mode capabilities of the preamplifier not be exceeded.

SOMETHING EVEN WORSE

As bad as power-frequency ground-loop interference can be, there is another kind of ground-loop interference that can pose far more severe problems. This more serious form of interference is coherent interference at the signal frequency produced when drive-signal current is allowed to flow through the braid of the signal cable. Figures 7 and 8 are provided to illustrate this problem and the steps that can be taken to prevent it. To begin with, Figure 7 shows an example where just about everything possible is done "wrong". The amplifier is operated single-ended. The ground connections at the experiment are made to the enclosure at many different spots, allowing currents to flow along the chassis, and worst of all, the drive signal currents are handled in a manner that allows them to flow through the braid of the signal cable.

In addition to providing the reference input signal to the lock-in amplifier, the drive signal can be presumed to be driving other system components as well. Depending on the nature of the experiment, these currents could range from very small to quite large, perhaps even amperes if the experiment involved driving a low impedance coil. Note that the various loads for the drive are represented by a single resistor returned to ground somewhere on the enclosure. Most of this drive signal current can be presumed

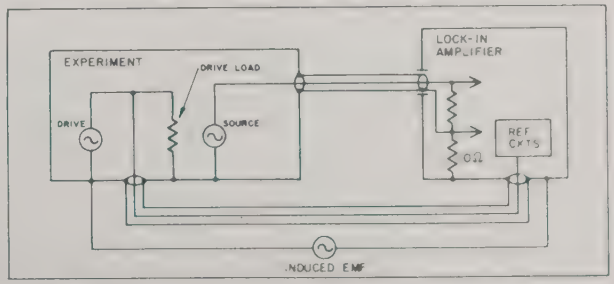

Figure 7. JUST ABOUT EVERYTHING "DONE WRONG"

to flow through the shield back to the drive signal source. However, a small but significant part of it will flow through the parallel path consisting of the braid of the signal cable, the ten ohm resistor, and the braid of the reference signal cable. The voltage drop of this current across the resistance of the signal-cable braid, even though attenuated by the ratio of the ten ohm resistor to the braid resistance, can constitute a serious source of interference at low signal levels, particularly since the interference is coherent, in phase with the signal, and directly adds to the signal. In short, the signal and the interference could well be indistinguishable, and if the interference is large enough to be significant, the measurement would be completely invalid.

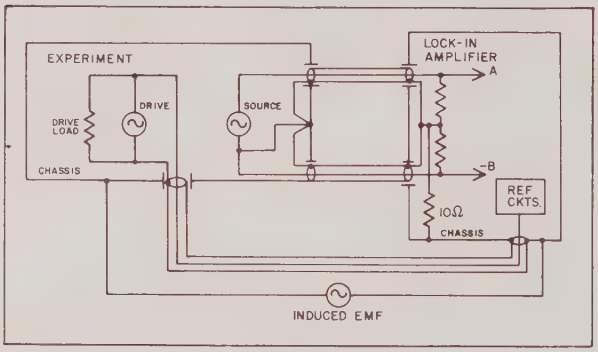

Figure 8. ERRORS CORRECTED

Figure 8 shows the steps that can be taken to circumvent this problem. All of the drive signal current is returned directly to the drive signal source, except for the small component that is applied to the lock-in amplifier by way of the Reference Input. Furthermore, the only coherent signal which can flow through the parallel path of the signal cable braid is a small portion of that allowed by the high input resistance of the Reference circuit. The use of differential operation assures that even this minute current will have no effect whatsoever. By operating in the manner indicated, one could use very high drive currents without concern that they might contaminate the signal of interest.

TRANSFORMERS

When working from a low-impedance source, noise considerations generally make it advisable to interpose a transformer between the source and the preamplifier. This transformer can either be external to the preamplifier, or, in the case of some preamplifiers such as the Models 116 and 119, internal to it. In any case, depending on how the transformer is connected, the input ground loop may be broken. Where this is the case, ground-loop interference is avoided. Figure 9 shows how a lock-in amplifier might be operated in conjunction with an external transformer such as the Model 190. Note that the input ground loop is broken by the transformer so that no ground-loop current can flow through the signal-cable braid. By connecting a transformer in this manner, single-ended operation with an extremely high immunity to ground-loop interference is possible.

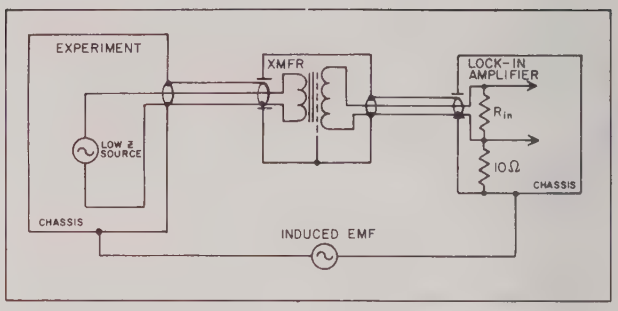

Figure 9. USE OF EXTERNAL SIGNAL TRANSFORMER

The situation is somewhat different, however, if the preamplifier were a Model 116 or 119 operated in the transformer mode. If single-ended operation were attempted, the situation would be as shown in Figure 10. The ground loop is not broken, although the benefit of the semi-floating ten ohm ground is maintained. The correct approach to using these preamplifiers with their internal transformers is to operate differentially as shown in Figure 11. Although the ground loop is not broken, the differential connection assures that there will be no response to any ground-loop current drops across the braid of the signal cables.

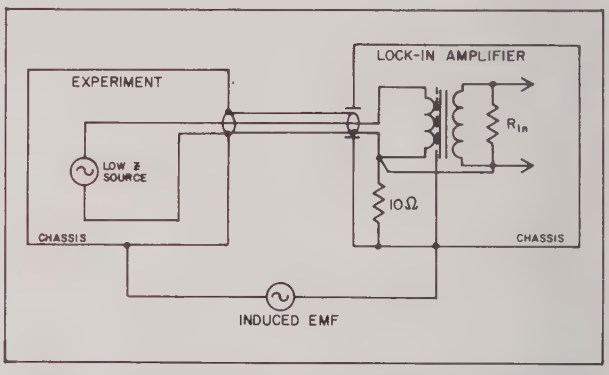

Figure 10. MODEL 116 OR 119 PREAMPLIFIER OPERATED IN TRANSFORMER MODE WITH SINGLE-ENDED CONNECTION

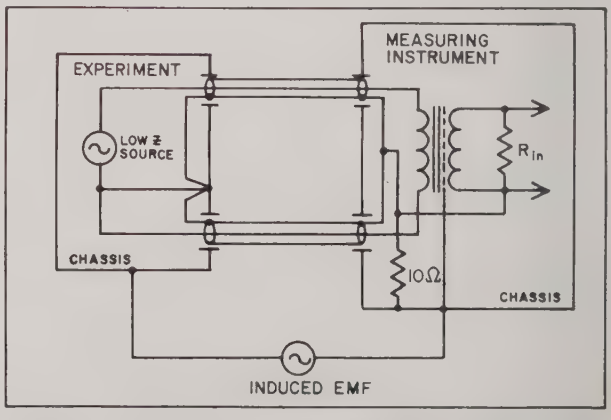

Figure 11. MODEL 116 OR 119 PREAMPLIFIER OPERATED IN TRANSFORMER MODE WITH DIFFERENTIAL CONNECTION

AND THEN THERE ARE CURRENT SOURCES

Lock-in amplifiers are commonly used to measure signals developed in current sources, such as photo-multiplier tubes. Figure 12 illustrates how the connection to a current source might be made. Generally speaking, ground-loop currents do not interfere with current source measurements. Because the source impedance is normally very much larger than R_{in}, E_{sbd} appears at both ends of R_{in} and so develops no input signal across R_{in}.

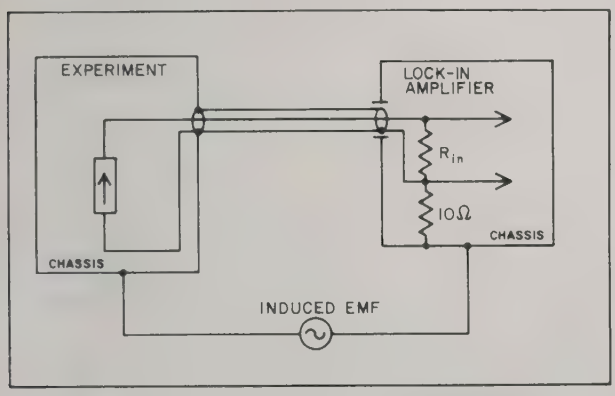

Figure 12. CURRENT SOURCE MEASUREMENT

It seldom happens that R_{in} is the ideal load when working from a current source. Usually there is a load resistor, smaller than R_{in} connected in parallel with R_{in}. When connecting this load resistor, it is important that it be connected at the preamplifier and not at the signal source. Best results are generally obtained by mounting the resistor in a shielded fixture fitted with BNC connectors so that the fixture can be connected directly to the preamplifier input. The signal cable is then connected to the other end of the fixture. The resistor should be positioned at the preamplifier end of the fixture so that the ground end of the resistor is as near the ground end of R_{in} as possible. If the load resistor is located at the source end of the signal cable, the braid drop potential can develop a ground-loop signal in R_{in} because of the parallel current path provided by the load resistor. With the load resistor close to R_{in}, there can be no potential difference between their respective ground returns, and no ground-loop signal will be developed in R_{in} and applied to the preamplifier.

Even though there is an inherent immunity to ground-loop interference in current-source applications, the situation is not quite so straightforward as indicated in the preceding paragraphs. It may happen that an operator will find he has severe ground-loop signal interference when working with a

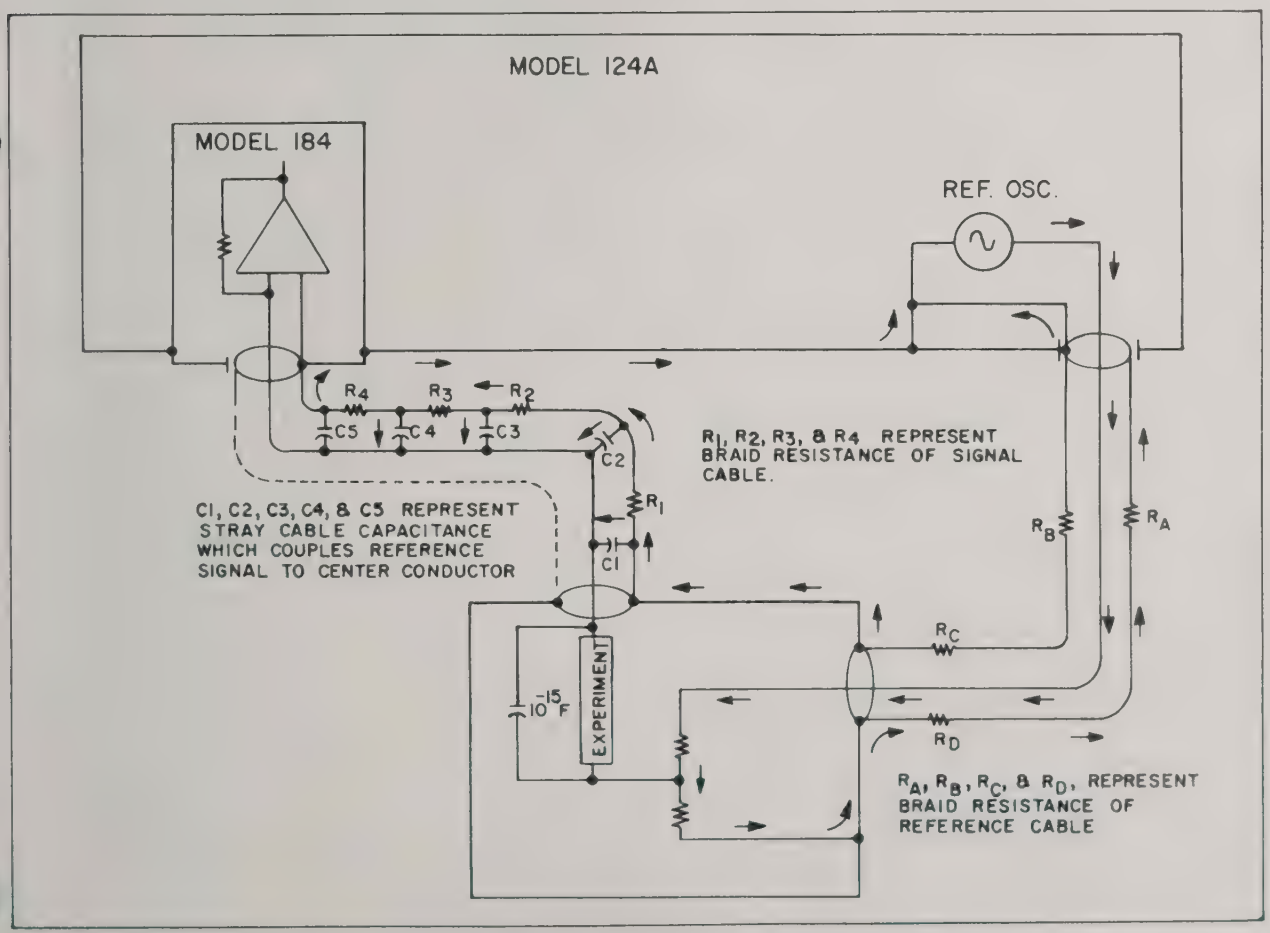

Figure 13. REFERENCE SIGNAL INTERFERENCE WITH MODEL 184 PREAMPLIFIER

current source. In light of the preceding discussion, such interference can be most perplexing, as there may seem to be no explanation for it. What happens is that the interference is getting in via the "back door" of the stray cable capacitance.

Figure 13 is an illustration of the Model 184 Photometric Preamplifier operated in conjunction with a Model 124A lock-in amplifier. Referring to the figure, note that the lock-in amplifier is assumed to be operating in the Internal Reference mode. Note also that the cable which extends from the lock-in amplifier reference output to the experiment, and that which extends from the experiment to the signal input of the Model 184, are both grounded at both ends, that is, the outer shell of each BNC connector is tied to the chassis. Although this may seem to be a reasonable way to operate, given the inherent immunity to ground-loop interference obtained in current source measurement applications, there are nevertheless possible problems due to the aforementioned "back-door" interference path, by which reference signal can get into the Model 184 input.

The solution to this problem is relatively simple once the mechanism of the reference signal coupling is understood. As shown in the figure, the reference oscillator can be thought of as a voltage generator inside the Model 124A. One side of the generator is returned to the Model 124A chassis, and the other side drives the center conductor of the cable which extends to the experiment. As the arrows indicate, the reference generator return path has two branches, one back through the braid of the reference cable and the other through the braid of the signal cable. Because the braids of these cables have a finite resistance, typically twenty or thirty milliohms, there is a voltage drop across the braid of both cables. This voltage, although it may be only a few microvolts, couples some reference signal current through the stray capacitance of the signal cable onto its center conductor. As a result, reference signal is applied to the input of the Model 184 and processed along with the information signal. This problem can be avoided completely by forcing all of the reference signal return current to flow back through the reference cable braid. To do that, one has only to break the connection between the signal cable and the chassis of the experiment. In other words, the signal cable should be grounded at the Model 184 Input end only.

It should be pointed out that there is nothing unique to operating the lock-in amplifier in the internal reference mode with regard to this problem. An analogous situation exists when the reference signal is derived from an external source. The solution is the same, that is, to see to it that the signal cable braid is grounded at the Model 184 input only.

Note that a 10^{-15} F capacitor is used in the diagram to represent the direct capacitive coupling from the reference circuits at the experiment to the signal line. This coupling could also constitute a source of interference. However, with reasonable construction techniques, this interference should prove much less a problem than the ground-loop interference discussed above.

THE MODEL 185, A SPECIAL CASE

The Model 185 differs from most of the other preamplifiers designed for use with the Model 124A in that it does not provide differential operation. This situation came about because the Model 185 represents the absolute state-of-the-art limit in low-noise preamplifier design. For reasons beyond the scope of this discussion, it is possible to achieve lower noise performance with a single-ended amplifier than with a differential amplifier, all things being equal.

Because single-ended amplifiers are more susceptible to ground-loop interference problems than are differential amplifiers, special attention was given to this problem in the Model 185 design. The solution selected was to provide the preamplifier with a semi-floating input ground of the type described earlier. However, whereas the other preamplifiers have ten-ohm grounds, the Model 185 has a 1000 ohm ground to achieve an exceptionally high immunity to ground-loop interference. The "tradeoff" one makes in using the semi-floating ground is one of gain accuracy. The gain of the instrument decreases inversely with the ratio of the 1000 ohms of the resistor to the 10 or 20 milliohms (typical) of the braid resistance. Even assuming a braid resistance of one ohm (unrealistically high), the Model 185 gain would decrease only 0.1%.

SOME FINAL POINTERS

For the purely technical physical consideration that it is, grounding has associated with it the most fantastic aura of folklore and black magic. Hopefully, this little paper has helped the reader to improve his basic grasp of ground-loop problems and the steps that can be taken to solve them. By applying the precepts discussed, the magic should seem to be a little less murky and more under control of the experimenter. Unfortunately, even when the problems are understood, the solutions are not always straightforward or simple. For example, if one has to use a preamplifier that is an inherently single-ended device without provision for floating the input or semi-floating the input ground, there may be relatively little one can do. Trying to convert such an instrument to floating or semi-floating operation will probably "disturb" it severely to say the least. These techniques must be designed into the preamplifier from the start. It may well be that the only solution is to "float" every chassis in the system. Of course, there may be the possibility of inserting a signal transformer or, better yet, of getting a different preamplifier. If the preamplifier in question can be operated differentially, use it that way, even when monitoring a "single-ended" source. However, if a signal transformer is being used in such a way as to break the input leg of the ground-loop system, then single-ended operation is perfectly feasible and even preferable.

In any case, the reader should now be prepared to approach his grounding problems with understanding, and not with just a vague urge to "strap everything together".

APPENDIX C

PREAMPLIFIER OVERLOAD DETECTION

One difference between the Model 124A and the earlier Model 124 is that the preamplifier overload sensing is done at a different point. In the Model 124, this sensing was done at the Selective Amplifier circuit board. In the Model 124A, the sensing is done at the preamplifier. As a consequence of this change, all plug-in preamplifiers supplied with the Model 124A differ from those supplied with the Model 124. The new-type preamplifiers can be identified by a small "x" located in the upper left-hand corner of the label at the rear of these instruments.

Proper preamplifier overload sensing is obtained if a new-type preamplifier is operated in conjunction with either a Model 124 or a Model 124A. Preamplifier output overload sensing is lost, however, if an old-type preamplifier is operated with a Model 124A. On two counts, this is not as serious as it may at first seem. First of all, the selective amplifier overloads first on most sensitivity ranges. The only ones where preamplifier overload can take place first are those for which the selective amplifier gain is X1 (least sensitive ranges; see Table V-1). Second, it is a simple matter to convert an old-type preamplifier to a new-type preamplifier. On the Models 116, 117, 118, 119, and 185, one need only connect two diodes (IN4009) to the blue-ribbon connector at the rear of the instrument. The diodes are connected inside the preamplifier as follows. One diode is connected from pin 11 (anode) to pin 14 (cathode). The other is connected from pin 10 (cathode) to pin 14 (anode). Nothing more is required.

In the case of the Model 184, the situation is a bit more complicated because an attenuator must be installed as well. The attenuator is formed by connecting a 4.02 kilohm resistor in series with a 2 kilohm resistor (both resistors 1%). The free end of the 4.02 kilohm resistor is connected to the junction of R216 and R217. The free end of the 2 kilohm resistor is connected to ground #1 (this is the ground bus to which the various bypass capacitors, C205, C210, C208, are returned). The diodes are connected to connector pins 10 and 11, the same as for the other preamplifiers. However, instead of returning the free ends of the diodes to pin 14, they are connected instead to the junction of the two attenuator resistors.

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